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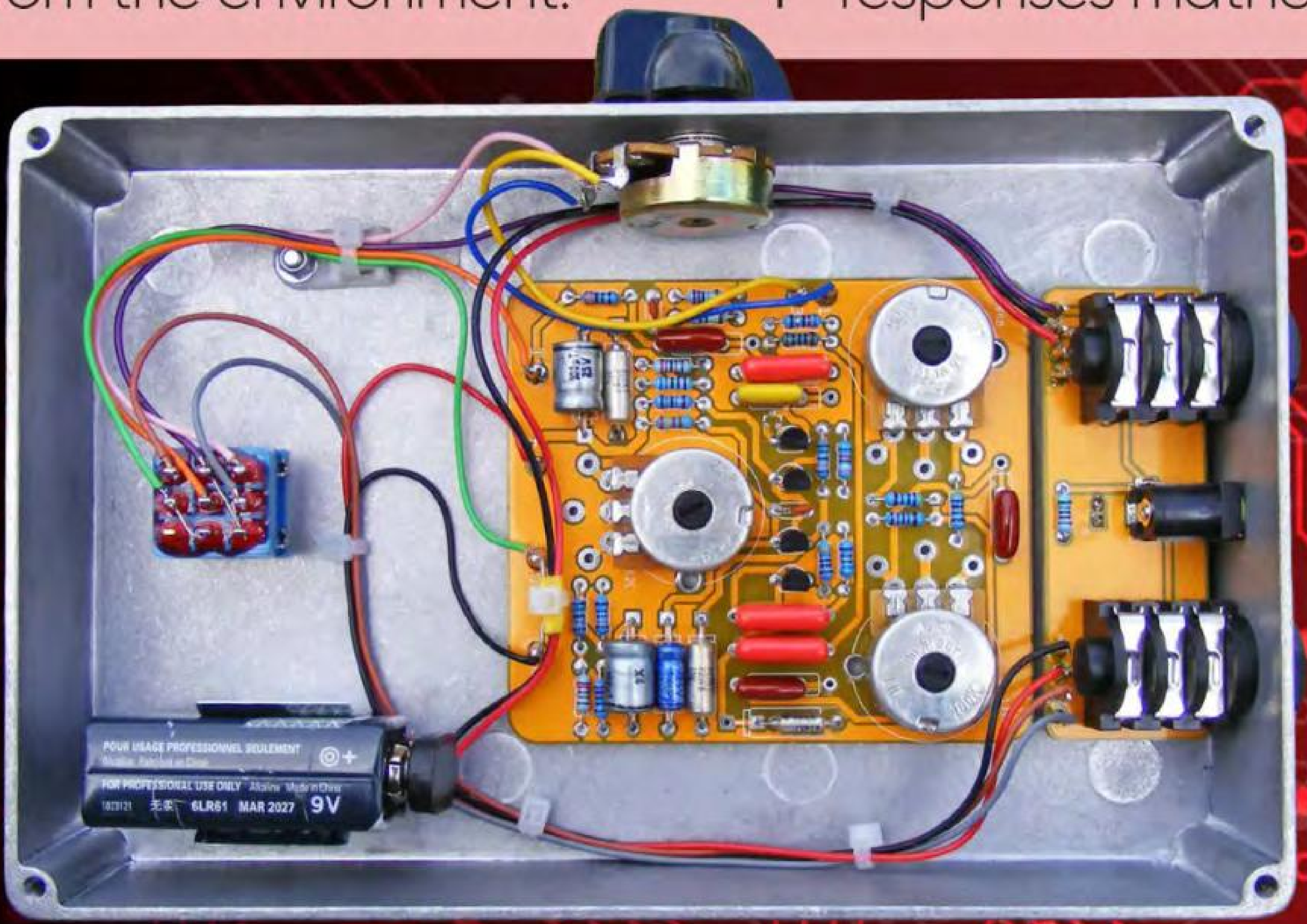
Part 3: radio transmitters and receivers.



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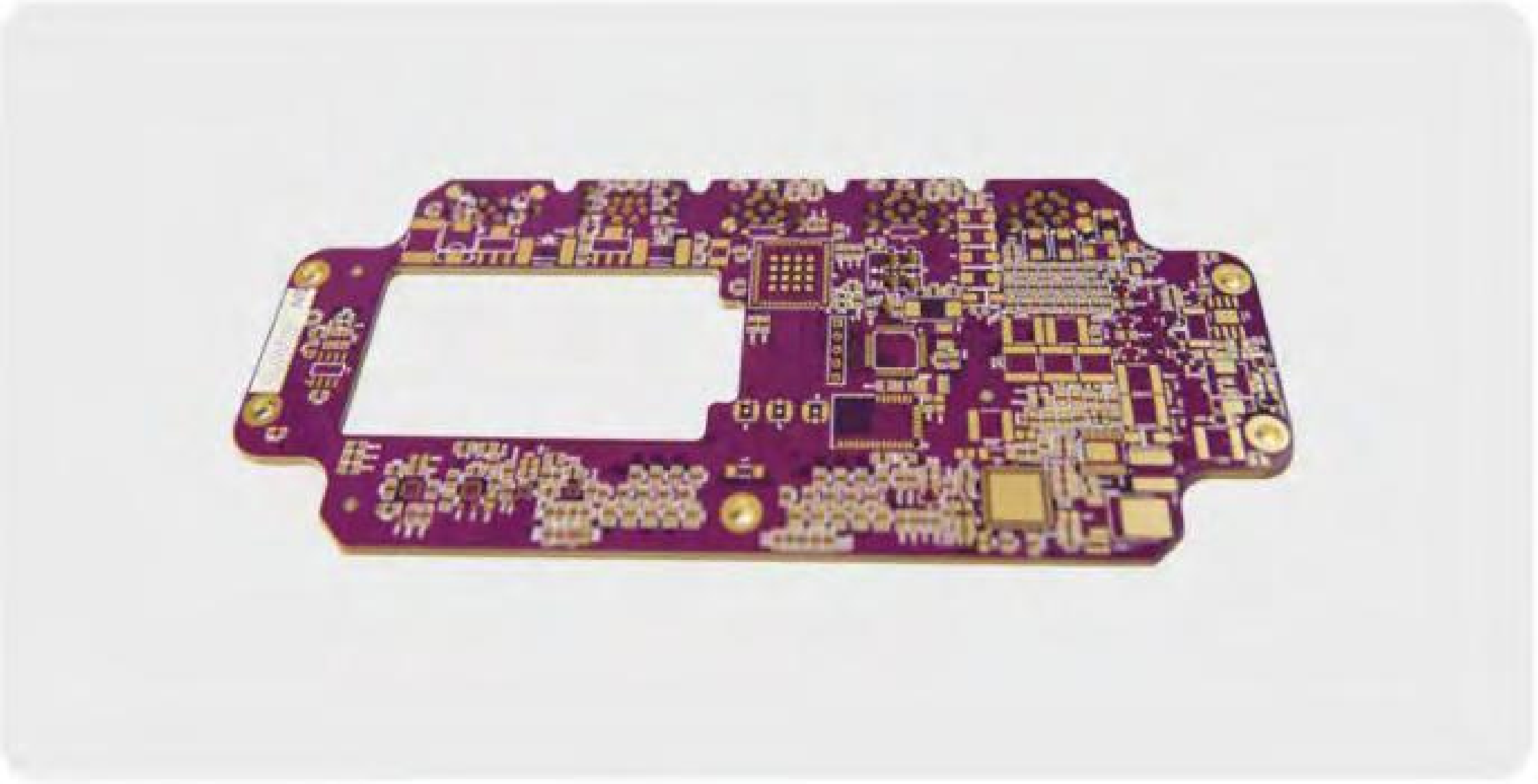

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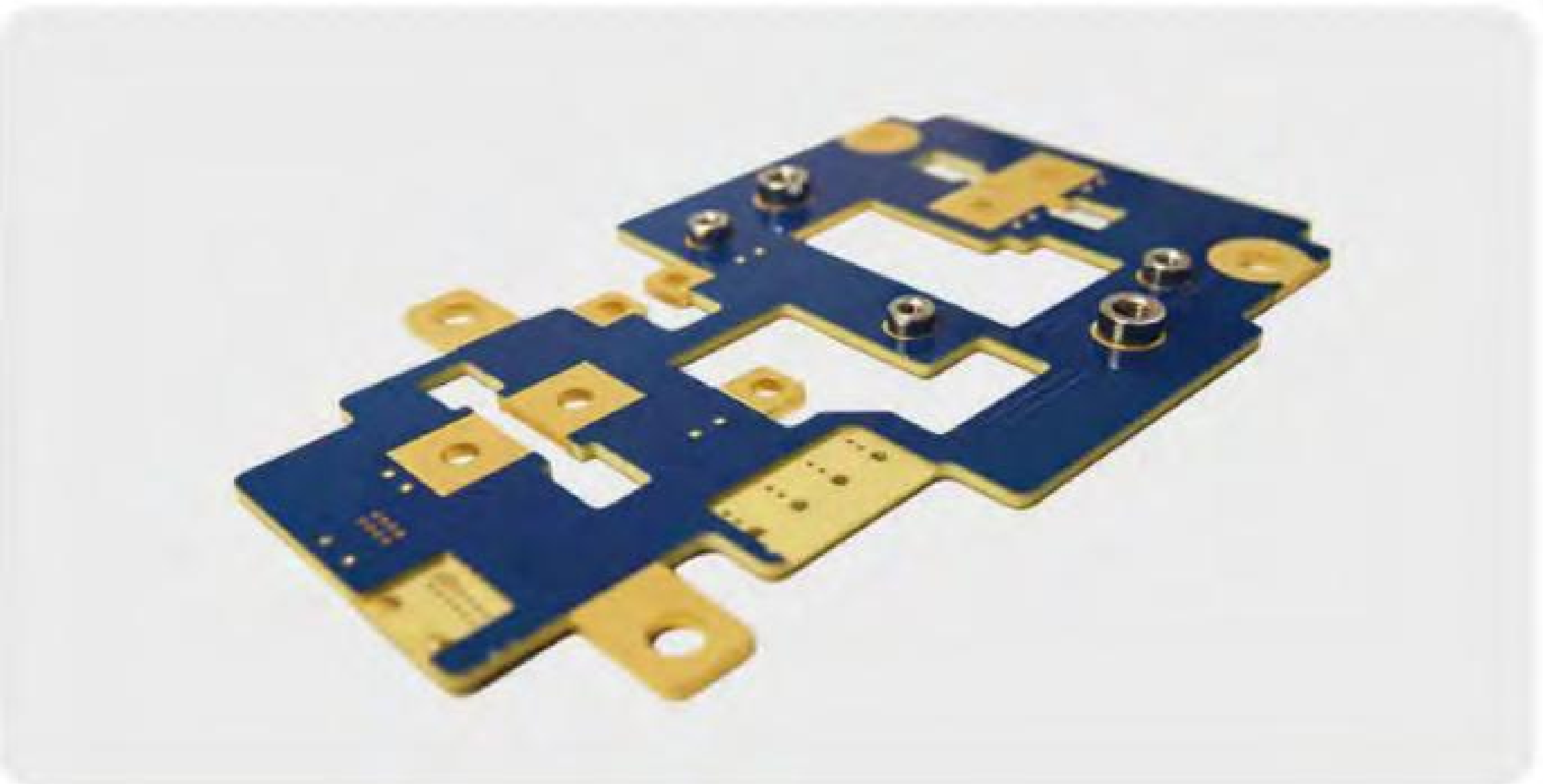

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Practical
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Volume 55. No. 01
January 2026
ISSN 2632 573X

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The February 2026 issue of *Practical Electronics* will be published on Tuesday, 6th of January, 2026 (see p80).

Renewals paid by PayPal, Visa and Mastercard are now handled via our website: **www.electronpublishing.com**

Practical Electronics

Volume 55. No. 01
January 2026
ISSN 2632 573X

# Editorial

**Editorial offices**
Electron Publishing (Australia)
**Tel** +61 2 9939 3295
**Email** pe@pemag.au
**Web** www.electronpublishing.com

**Address mail to:**
Electron Publishing (Australia)
PO Box 194, Matraville NSW 2036
Australia

**Advertising enquiries**
+61 2 9939 3295 pe@pemag.au

| | |
|---|---|
| **Editor** | Nicholas Vinen |
| **Publisher** | Nicholas Vinen |
| **Digital subscriptions** | Stewart Kearn **Tel 07918 614662** |
| **Online Editor** | Alan Winstanley |
| **Web Systems** | Kris Thain |
| **Production** | Bao Smith |
| **Technical staff** | Tim Blythman, John Clarke |

**Print subscriptions**
Practical Electronics Subscriptions
PO Box 6337
Bournemouth BH1 9EH
United Kingdom
**Tel** 01202 087631
**Email** pesubs@selectps.com

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## Hardware requiring an App is a red flag

Lately, I've noticed more consumer gear (routers, printers etc) requiring a smartphone App to control or set it up. I plan to avoid buying such gear in future for a few reasons, the primary one being the obvious: it will have a limited lifespan.

For example, I bought a TP-Link Deco mesh Wi-Fi system for my home office because a single Wi-Fi router, even a high-end one, can't penetrate the double brick walls very well. This system works well, but it's fully controlled via a smartphone App. What's stopping TP-Link from withdrawing the App and preventing me from changing its configuration, or even remotely disabling the whole system? Nothing, really. And that is not without precedent, either.

Had I known that this system was App-configured when I bought it, I probably would have chosen something else. Next time, I will certainly do more research and buy a system that doesn't rely on an App. I don't mind having an optional App, but I want another configuration method for the inevitable day when the App is no longer available.

Of course, companies love this because it's yet another form of planned obsolescence. iFixit (a right-to-repair-friendly company) has a good article on this unfriendly practice on their website here:

https://www.ifixit.com/News/97135/companies-keep-bricking-products-who-will-hold-them-accountable (https://pemag.au/link/ac9d).

**Nicholas Vinen, Electron Publishing (Australia)***
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TREBLE
AMP
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BASS
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Jake Rothman's "Pre-driver" guitar preamp/ overdrive pedal, described on pages 50-55 (no App needed!).

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# Teach-In 2026

### by Mike Tooley

## World of Wireless – An Introduction to Radio and Wireless Technology

## Series 12, part 3: transmitters and receivers

**In the previous instalment, we** presented an overview of RF circuits and their associated components. The discussion began with a simple VHF signal source, followed by an analysis of frequency-selective resonant circuits and RF filters.

We then examined the application of quartz resonators as replacements for conventional L-C tuned circuits, highlighting their effectiveness in constructing filters with superior selectivity. Additionally, we provided a list of representative semiconductors suited for RF applications, several of which will feature in forthcoming practical exercises throughout this series.

Our companion Hands-On project showcased an enhanced portable medium wave AM receiver.

This month, our focus shifts to fundamental radio communication systems. We discuss the necessity for modulation and demodulation, and introduce the primary methods—AM and FM—following an initial overview of Morse code as a basic communication technique.

Our Hands-On projects involve building a crystal-controlled reference oscillator and a separate variable-frequency oscillator (VFO). These two handy modules will be used in conjunction with several of our subsequent Hands-On projects.

### A simple radio communication system

Let's start with a very simple radio communication system comprising a transmitter (TX) and receiver (RX) in which communication is achieved by simply switching (or 'keying') the radio frequency signal on and off (see Fig.3.1).

Keying can be achieved by interrupting the supply to the power amplifier stage, or even the oscillator stage. However, keying the oscillator stage usually results in poor frequency stability, so it is normally applied within a driver stage operating at a modest power level.

The receiver consists of nothing more than a radio-frequency amplifier (which provides gain and selectivity) followed by a detector and an audio amplifier. The detector stage mixes a locally generated radio-frequency signal produced by the beat frequency oscillator (BFO) with the incoming signal to produce a signal that lies within the audio frequency range (typically between 300Hz and 3.4kHz).

As an example, assume that the incoming signal is at a frequency, $f_{RF}$, of 100kHz and that the BFO is producing a signal, $f_{BFO}$, at 99kHz. A signal at the difference between these two frequencies (1kHz), $f_{AF}$, will appear at the output of the detector stage. This will then be amplified by the audio stage before being fed to the loudspeaker. The relationship between these three frequencies is $f_{BFO} = f_{RF} \pm f_{AF}$.

Note that the BFO can operate above or below the incoming signal frequency by an amount equal to the desired beat frequency (the audible signal resulting from the 'beating' of the two frequencies).

### Morse code

Morse code transmissions are

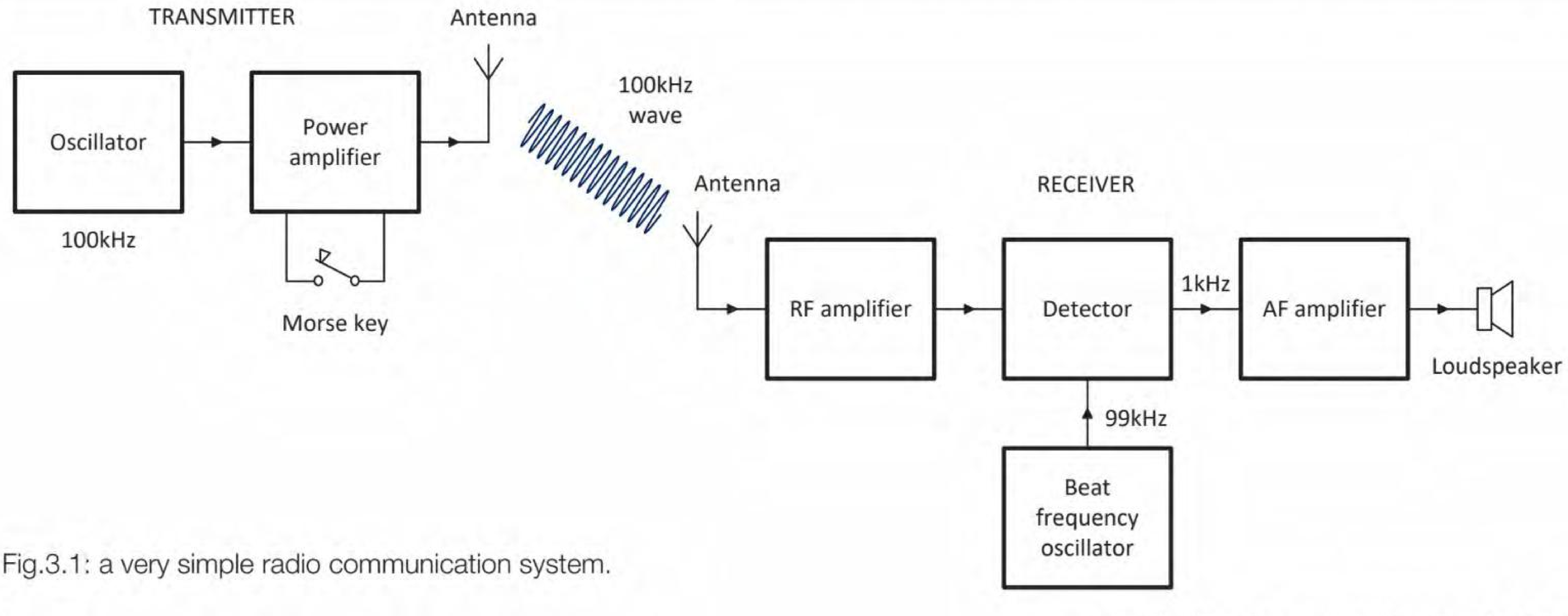


Fig.3.1: a very simple radio communication system.

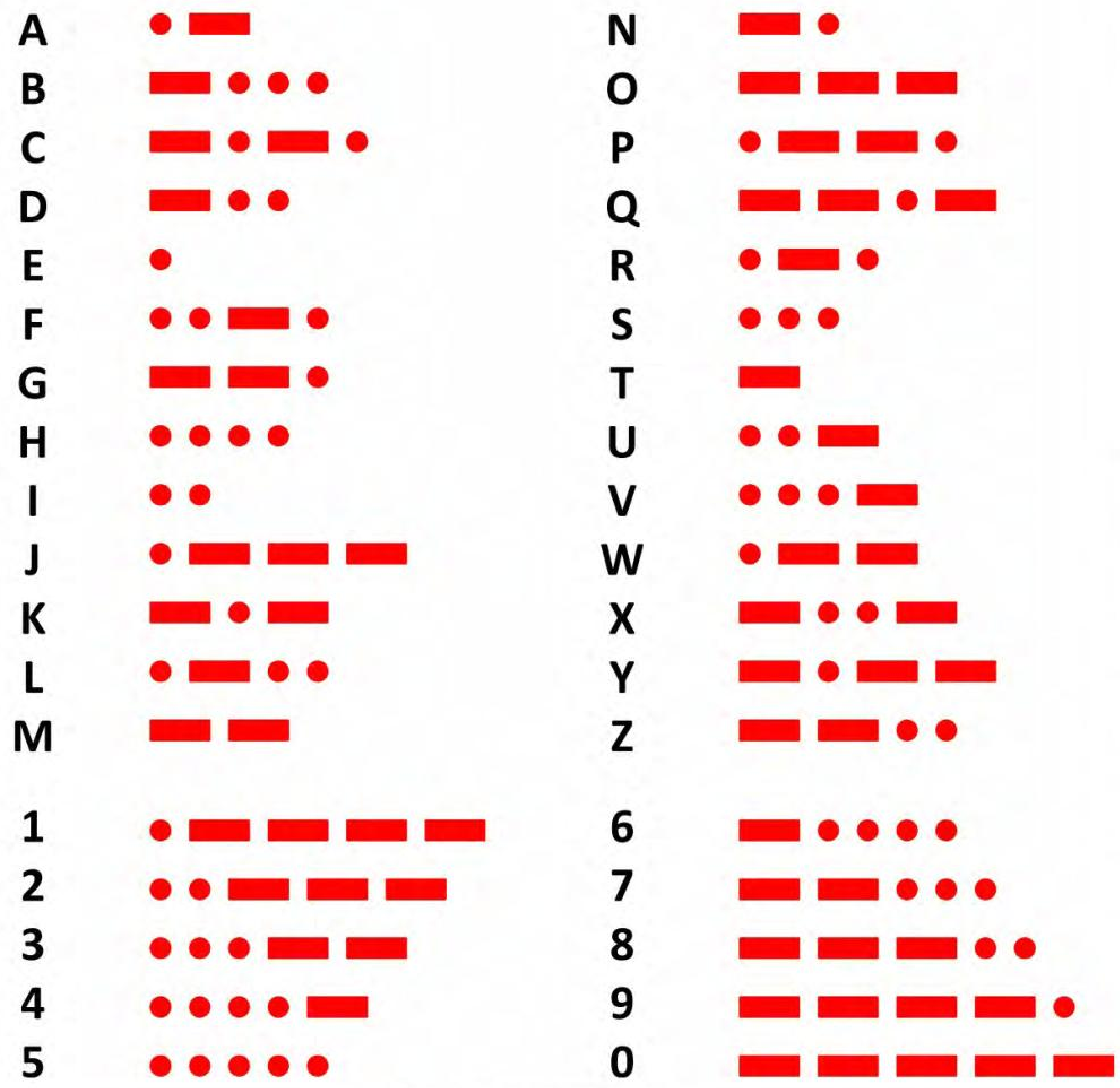


Fig.3.2: Morse code.

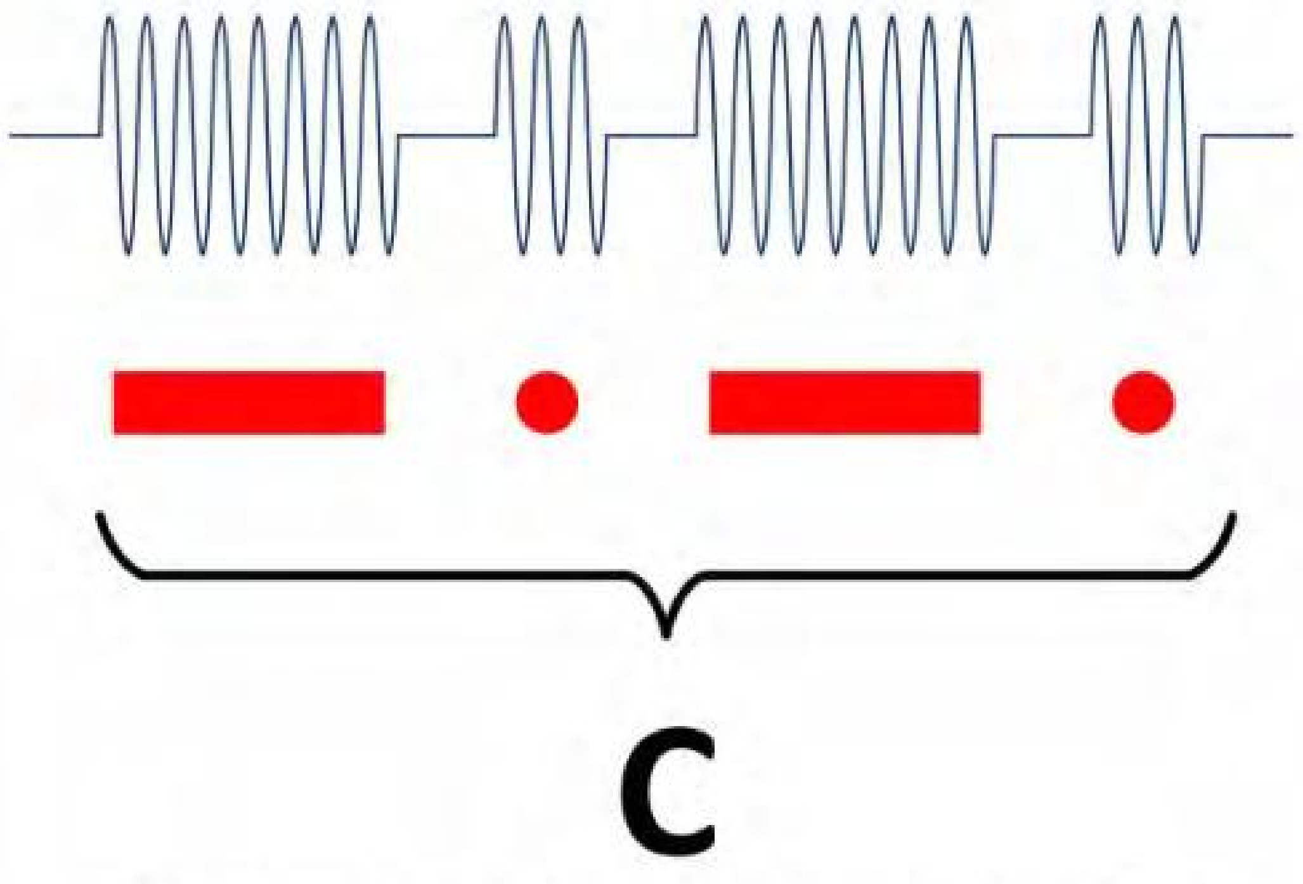


Fig.3.3: the letter "C" sent using an interrupted RF carrier wave.

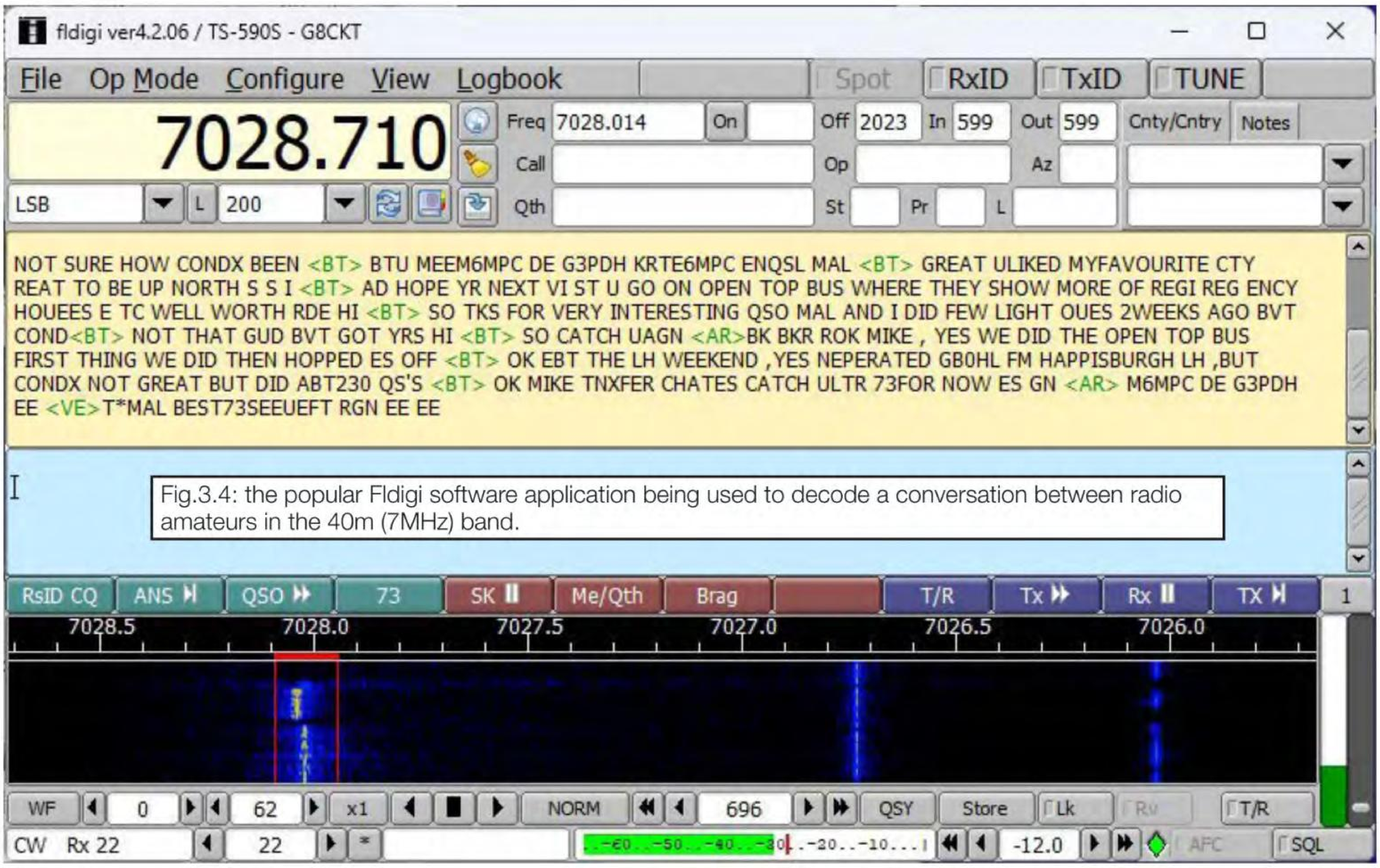
often referred to as continuous wave (CW), but this term can be a little misleading as the wave is not actually continuous but periodically interrupted to send each character.

Transmitters and receivers for CW operation are extremely simple but nevertheless can be very efficient. This makes them particularly useful for disaster and emergency communication, or any situation that requires optimum use of low-power equipment. Signals are transmitted using the code invented by Samuel Morse (see Figs.3.2 & 3.3).

Morse Code may seem an outdated form of communication, but it is still used in several practical applications, including aeronautical beacons, amateur radio and emergency communications.

Although speeds of around 15-20 words per minute are typical, a skilled operator can send code using a straight key at more than 25 words per minute, and up to 40 words per minute using a paddle and electronic keyer. In addition, software is available that supports both sending and receiving using a computer.

Fig.3.4 shows the immensely popular Fldigi application being used to decode a Morse code exchange between radio amateurs in the 40-metre (7MHz) band. It's important to note that, unlike the ear of an experienced Morse operator, decoding software can often struggle in the presence of noise, fading and interference!



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HOUEES E TC WELL WORTH RDE HI <BT> SO TKS FOR VERY INTERESTING QSO MAL AND I DID FEW LIGHT OUES 2WEEKS AGO BVT
COND<BT> NOT THAT GUD BVT GOT YRS HI <BT> SO CATCH UAGN <AR>BK BKR ROK MIKE , YES WE DID THE OPEN TOP BUS
FIRST THING WE DID THEN HOPPED ES OFF <BT> OK EBT THE LH WEEKEND ,YES NEPERATED GB0HL FM HAPPISBURGH LH ,BUT
CONDX NOT GREAT BUT DID ABT230 QS'S <BT> OK MIKE TNXFER CHATES CATCH ULTR 73FOR NOW ES GN <AR> M6MPC DE G3PDH
EE <VE>T*MAL BEST73SEEUEFT RGN EE EE
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WF 0  62  696  CW Rx 22  22  -12.0

Fig.3.4: the popular Fldigi software application being used to decode a conversation between radio amateurs in the 40m (7MHz) band.

## Modulation

To convey information using a radio-frequency carrier, the signal information must be superimposed or 'modulated' onto the carrier wave. Modulation is the name given to the process of changing a particular property of the carrier wave in sympathy with the instantaneous voltage (or current) of the signal to be conveyed.

The most common modulation methods are amplitude modulation (AM) and frequency modulation (FM). In the former case, the carrier amplitude (its peak voltage) varies according to the voltage, at any instant, of the modulating signal. In the latter case, the carrier frequency is varied in accordance with the voltage, at any instant, of the modulating signal.

Fig.3.5 shows the effect of amplitude and frequency modulating a sinusoidal carrier (the modulating signal is in this case also sinusoidal). In practice, many more cycles of the RF carrier would occur in the span of one cycle of the modulating signal.

The modulating of a carrier is undertaken by a modulator circuit. The input and output waveforms for amplitude and frequency modulator circuits are shown respectively in Figs.3.6(a) & 3.6(b).

Common examples of the use of amplitude modulation (AM) include long-wave, medium-wave and short-wave broadcasting, as well as VHF aircraft communication. Common examples of the use of frequency modulation (FM) include FM broadcasting, radio microphones and low-cost handheld walkie-talkies.

## Demodulation

Demodulation is the reverse of modulation and is how the signal information is recovered from the modulated carrier. Demodulation is achieved by means of a demodulator (sometimes also called a 'detector'). The output of a demodulator consists of a reconstructed version of the original signal information present at the input of the modulator stage at the transmitter.

The input and output waveforms for amplitude and frequency demodulator circuits are shown in Figs.3.7(a) & 3.7(b), respectively.

## AM transmitters

Fig.3.8 shows the block diagram of a simple AM transmitter. An accurate and stable RF oscillator generates the radio-frequency carrier signal. The output of this stage is then amplified and passed to a modulated RF power amplifier stage. The inclusion of an amplifier between the RF oscillator and the modulated stage also helps to improve frequency stability.

The low-level signal from the microphone is amplified using an AF amplifier before it is passed to an AF power amplifier. The output of the power amplifier is then fed as the supply to the modulated RF power amplifier stage. Increasing and reducing the supply to this stage is instrumental in increasing and reducing the amplitude of its RF output signal.

The modulated RF signal then passes through an antenna coupling unit. This matches the RF power amplifier's output to the antenna

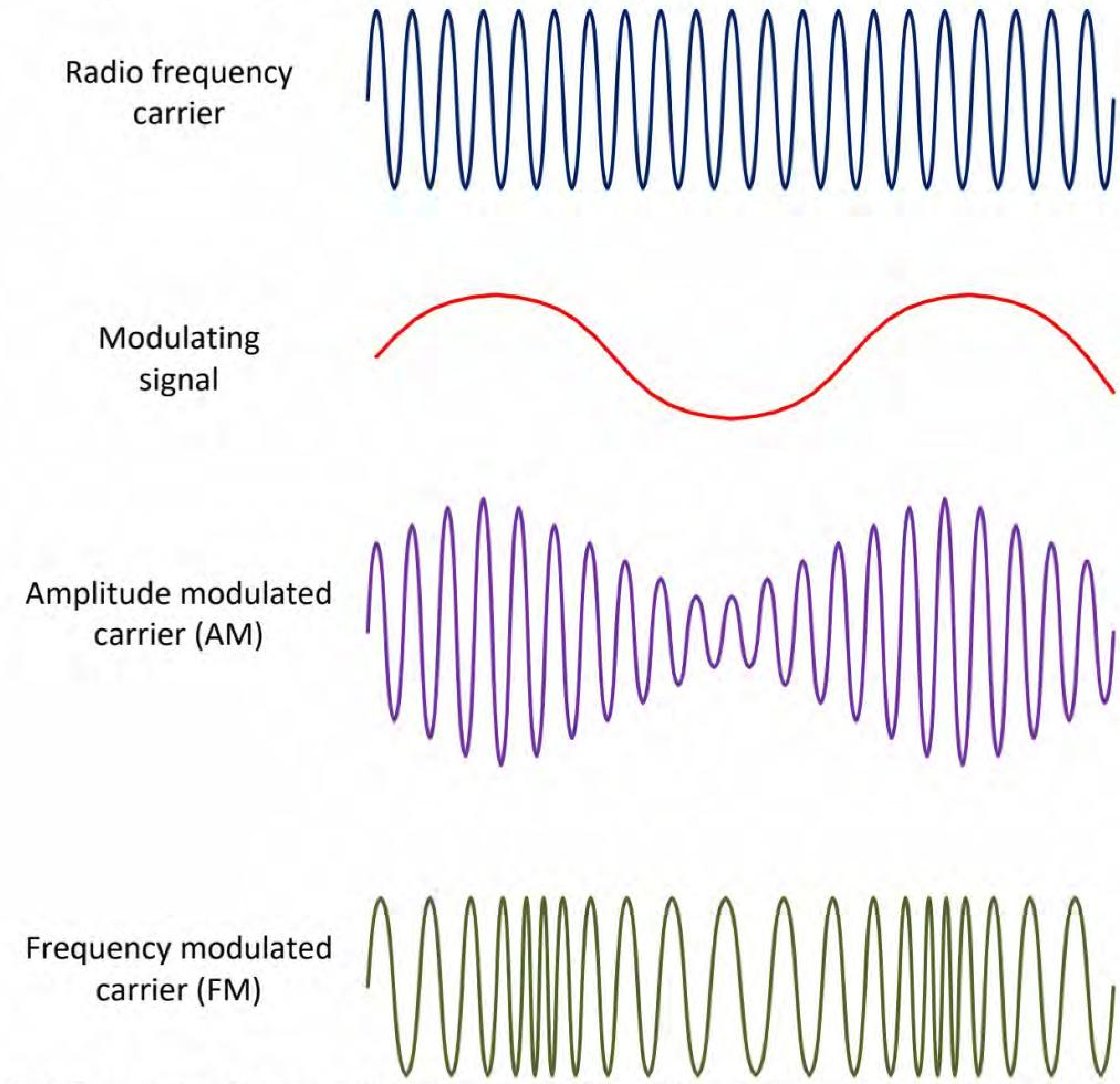

Radio frequency carrier
Modulating signal
Amplitude modulated carrier (AM)
Frequency modulated carrier (FM)

Fig.3.5: note the differences between amplitude and frequency modulation.

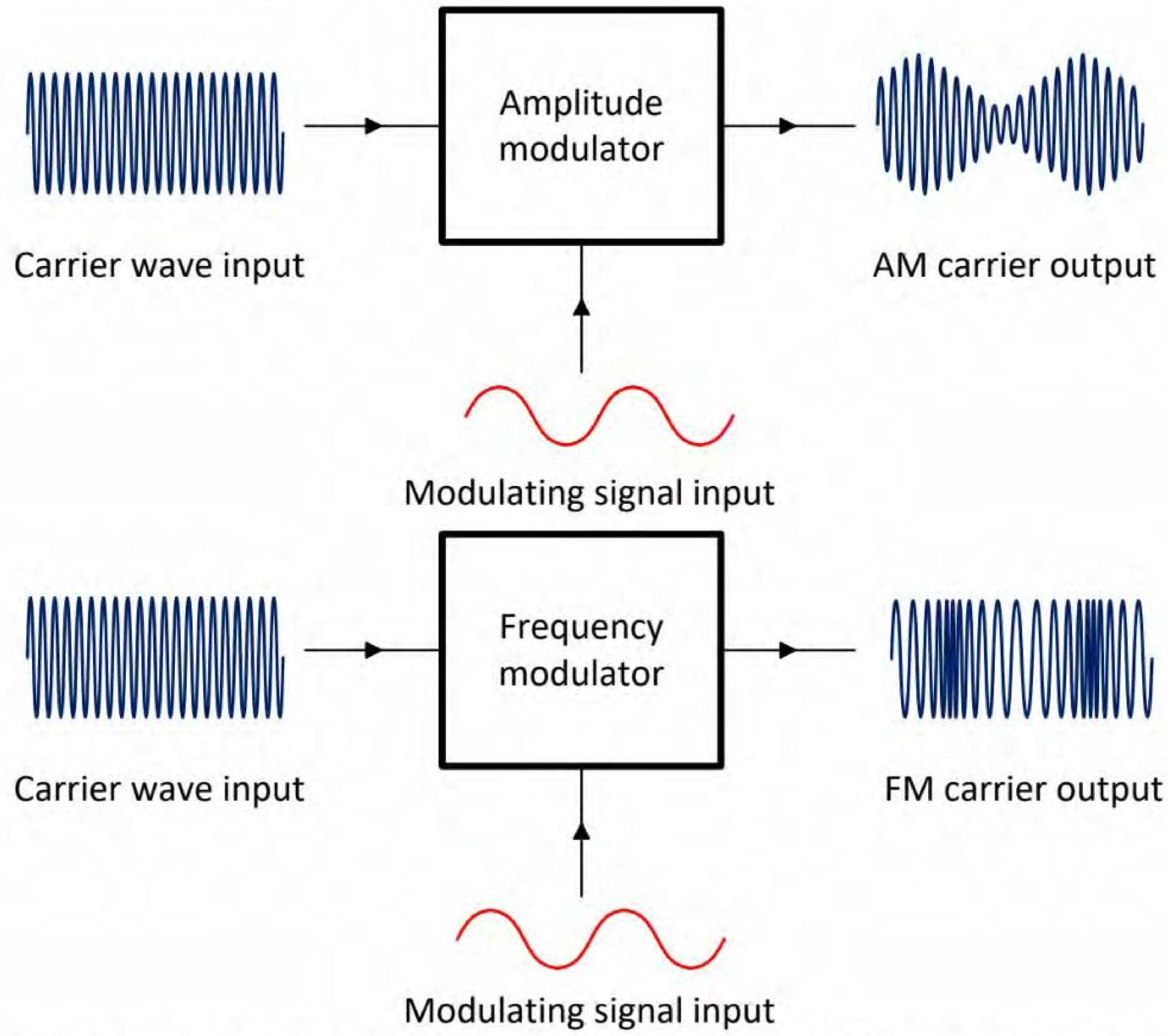


Fig.3.6: a simplified way to look at the circuitry required for AM (upper) & FM (lower).

and helps to reduce the level of any unwanted harmonic components that may be present. The AM transmitter shown in Fig.3.8 employs high-level modulation, where the modulating signal is applied to the final RF power amplifier stage.

An alternative to high-level modulation of the carrier wave is shown in Fig.3.9. In this arrangement, the modulation is applied to a low-power RF amplifier stage. The amplitude-modulated signal is then further amplified by the final RF power amplifier stage.

To prevent distortion of the modulated waveform, this final stage must operate in linear mode (the output waveform must be a faithful replica of the input waveform). Low-level modulation avoids the need for an AF power amplifier, but the RF power amplifier must operate in a linear mode. If this isn't the case, the transmitted signal will be distorted and unwanted signal components will be generated.

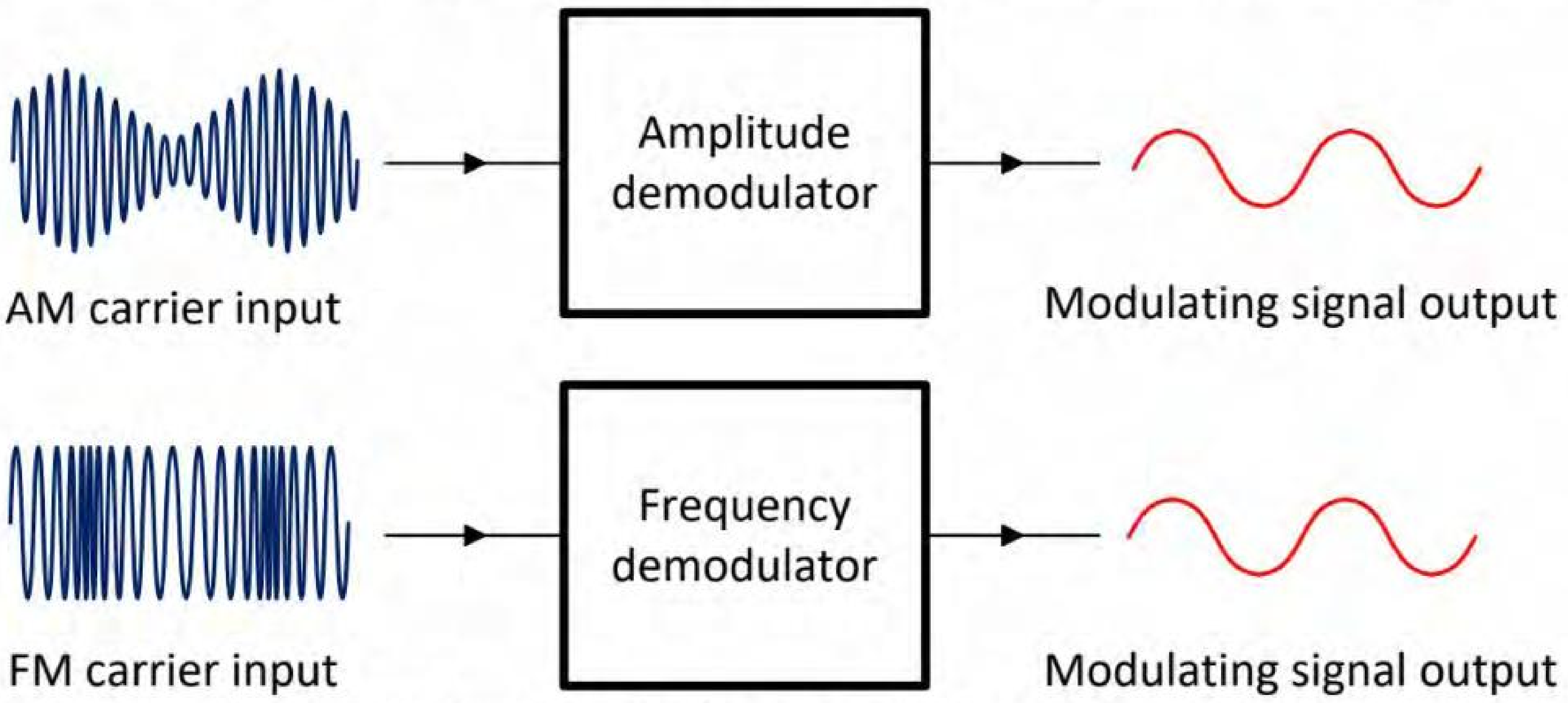

AM carrier input
Amplitude demodulator
Modulating signal output
FM carrier input
Frequency demodulator
Modulating signal output

Fig.3.7: simplified amplitude demodulation (upper) & frequency demodulation (lower).

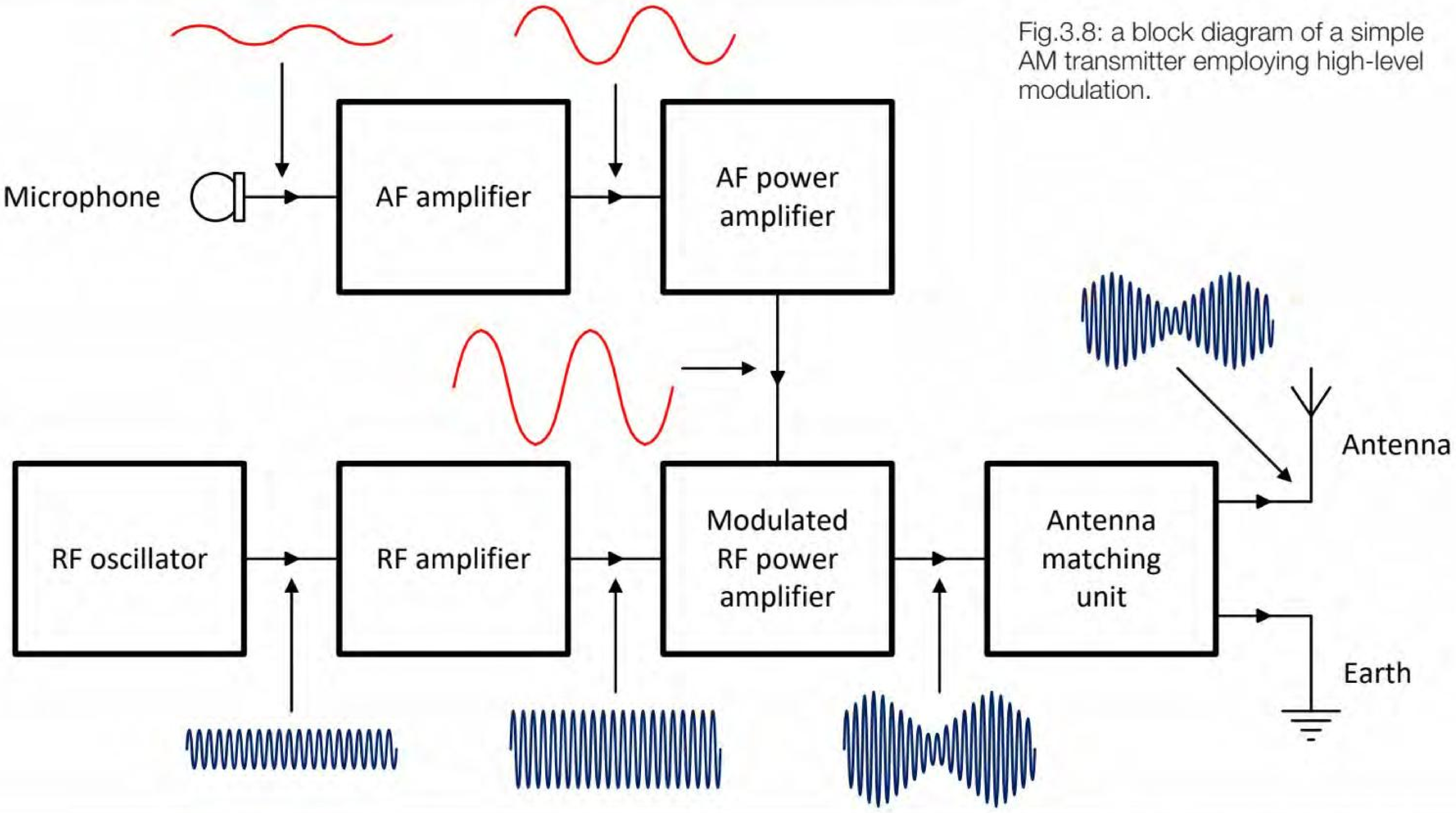

Microphone
AF amplifier
AF power amplifier
RF oscillator
RF amplifier
Modulated RF power amplifier
Antenna matching unit
Antenna
Earth

Fig.3.8: a block diagram of a simple AM transmitter employing high-level modulation.

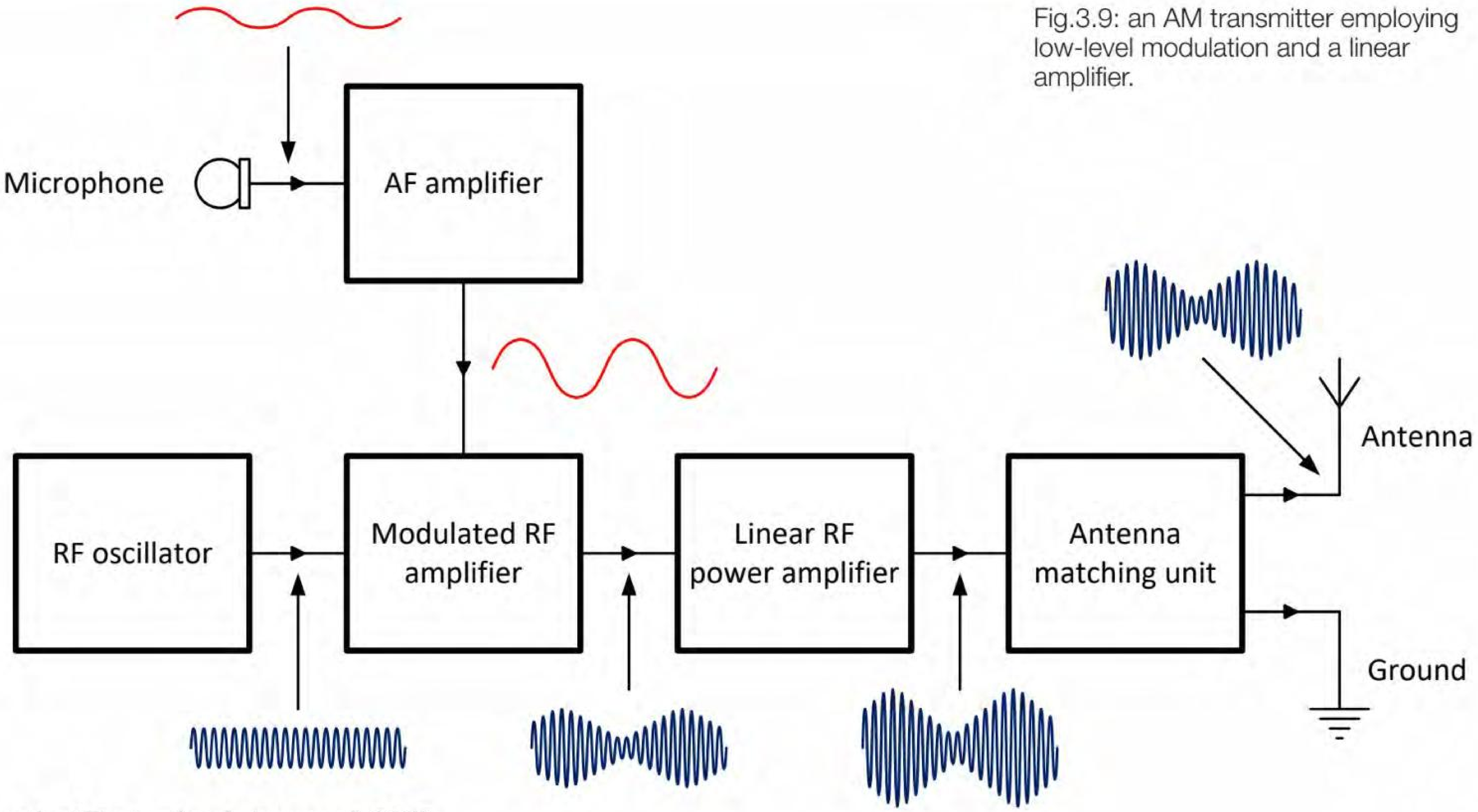
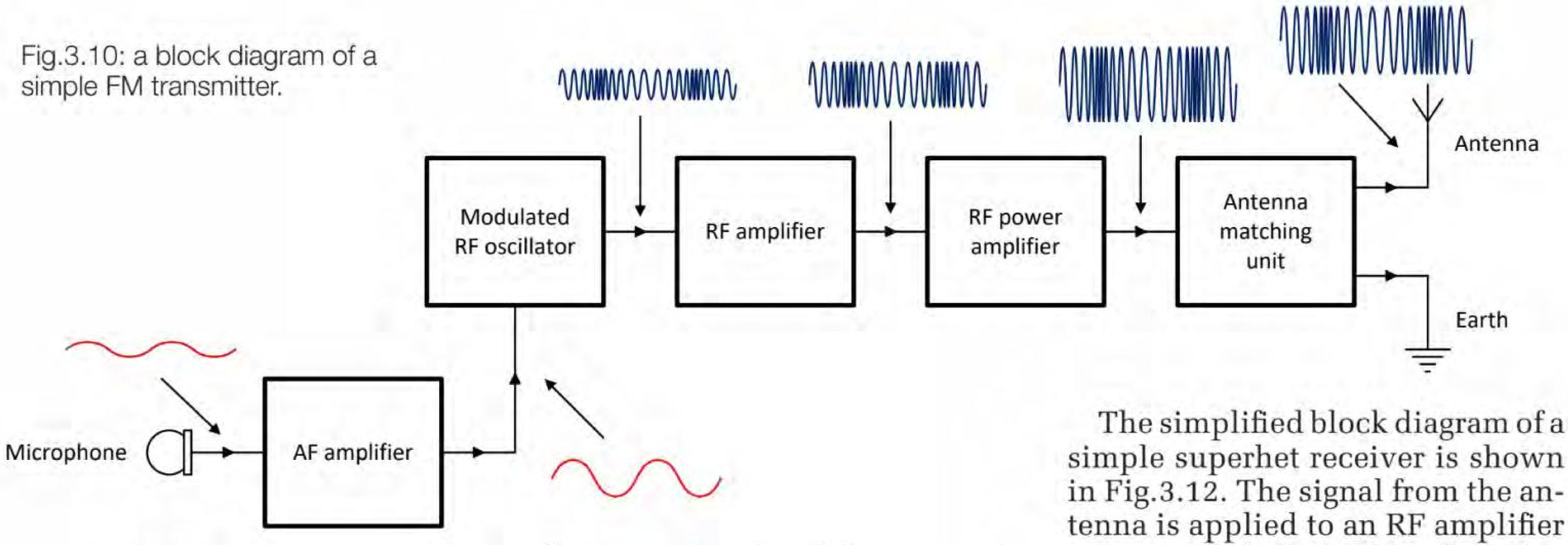

Microphone
AF amplifier
RF oscillator
Modulated RF amplifier
Linear RF power amplifier
Antenna matching unit
Antenna
Ground

Fig.3.9: an AM transmitter employing low-level modulation and a linear amplifier.

Fig.3.10: a block diagram of a simple FM transmitter.



### FM transmitters

Fig.3.10 shows the block diagram of a simple FM transmitter. Once again, an accurate and stable RF oscillator generates the radio frequency carrier signal. As with the AM transmitter, the output of this stage is amplified and passed to an RF power amplifier stage. Here again, the inclusion of an amplifier between the RF oscillator and the RF power stage helps to improve frequency stability.

The low-level signal from the microphone is amplified using an AF amplifier before it is passed to a variable reactance element (eg, a variable capacitance diode) within the RF oscillator tuned circuit. The application of the AF signal to the variable reactance element causes the frequency of the RF oscillator to increase and decrease in sympathy with the AF signal.

As with the AM transmitter, the final RF signal from the power amplifier is passed through an antenna coupling unit that matches the antenna to the RF power amplifier and may also help reduce the level of any unwanted harmonic components.

### TRF receivers

Tuned radio frequency (TRF) receivers provide a means of receiving local signals using minimal circuitry, and as such, were among the earliest receivers developed. The simplified block diagram of a TRF receiver is shown in Fig.3.11.

The signal from the antenna is applied to an RF amplifier stage. This stage provides a moderate amount of gain at the signal frequency. It also provides selectivity by incorporating one or more tuned circuits at the signal frequency. This helps the receiver reject signals that may be present on adjacent channels.

The output of the RF amplifier stage is applied to the demodulator. This stage recovers the audio frequency signal from the modulated RF signal. The demodulator stage may also incorporate a tuned circuit to further improve the selectivity of the receiver.

The output of the demodulator stage is fed to the input of the AF amplifier stage. This stage increases the level of the audio signal from the demodulator so it is sufficient to drive a loudspeaker. Unfortunately, TRF receivers have several limitations regarding sensitivity and selectivity, making this type of radio receiver generally unsuitable for commercial radio equipment.

### Superheterodyne receivers

Superhet receivers provide both improved sensitivity (the ability to receive weak signals) and improved selectivity (the ability to discriminate signals on adjacent channels) compared to TRF receivers.

Superhet receivers are based on the supersonic heterodyne principle, where the wanted input signal is converted to a fixed intermediate frequency (IF) at which most of the gain and selectivity is applied. The intermediate frequency chosen is generally 455kHz or 1.6MHz for AM receivers and 10.7MHz for communications and FM receivers.

The simplified block diagram of a simple superhet receiver is shown in Fig.3.12. The signal from the antenna is applied to an RF amplifier stage. As with the TRF receiver, this stage provides a moderate amount of gain at the signal frequency. The stage also provides selectivity by incorporating one or more tuned circuits at the signal frequency.

The output of the RF amplifier stage is applied to the mixer stage. This stage combines the RF signal with the signal derived from the local oscillator (LO) stage to produce a signal at the intermediate frequency (IF).

It is worth noting that the output signal produced by the mixer contains several components, including the sum and difference of the signal and local oscillator frequencies as well as the original signals plus harmonic components.

The wanted signal (ie, at the intermediate frequency) passes (usually via a bandpass filter—see Part 2) to the IF amplifier stage. This stage provides amplification as well as a high degree of selectivity.

The output of the IF amplifier stage is fed to the demodulator stage. As with the TRF receiver, this stage is used to recover the audio-frequency signal from the modulated RF signal. Finally, the AF signal from the demodulator stage is fed to the AF amplifier. As before, this stage increases the level of the audio signal from the demodulator so that it is sufficient to drive a loudspeaker.

To cope with a wide variation in input signal amplitude, superhet receivers invariably incorporate some form of automatic gain control (AGC). In most circuits, the DC level

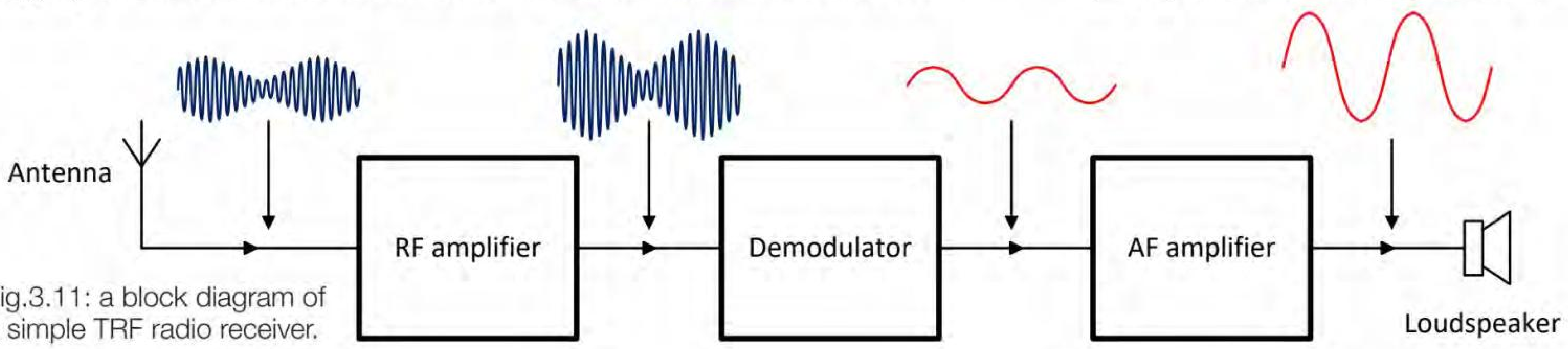


Fig.3.11: a block diagram of a simple TRF radio receiver.

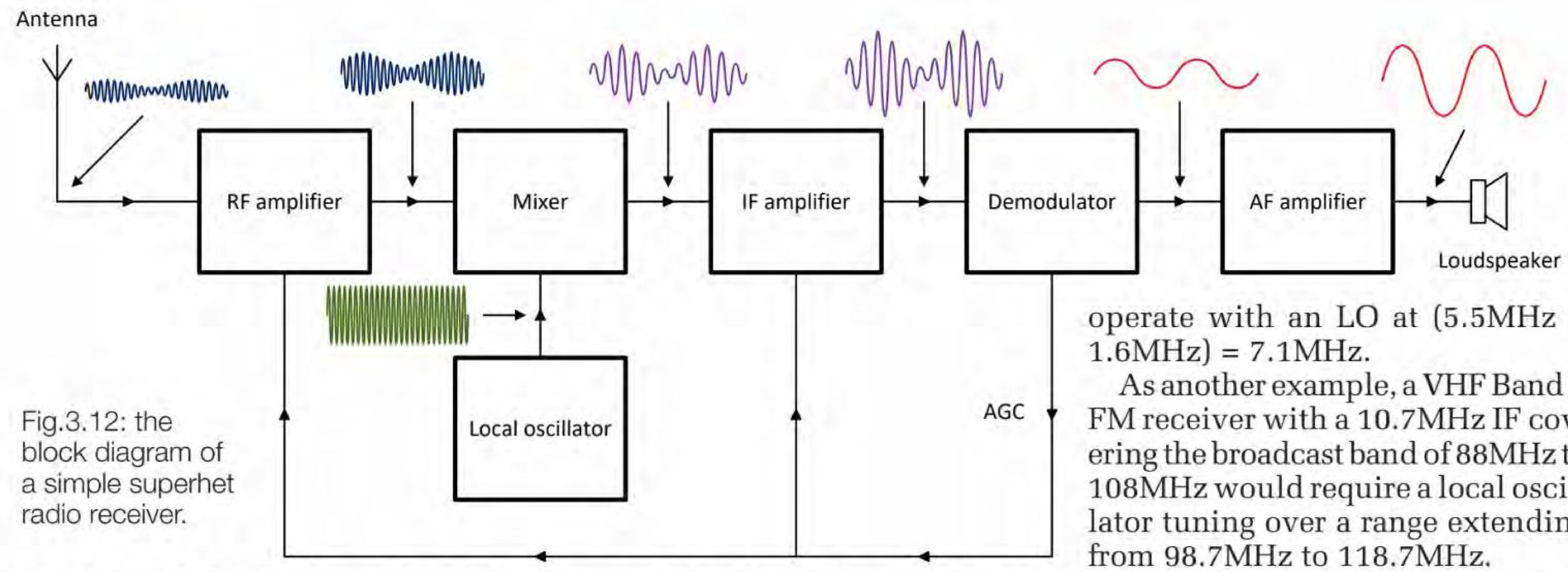


Fig.3.12: the block diagram of a simple superhet radio receiver.

from the AM demodulator is used to control the gain of the IF and RF amplifier stages.

As the signal level increases, the DC level from the demodulator stage increases, reducing the gain of both the RF and IF amplifiers. This provides a similar audio output level for both weak and strong stations.

### Frequency conversion

The superhet receiver's intermediate frequency is the difference between the signal frequency and the local oscillator frequency. The required local oscillator frequency can be calculated from the relationship $f_{LO} = f_{RF} \pm f_{IF}$.

In most cases, to simplify tuning arrangements and reduce the impact of 'image' channels, the local oscillator (LO) usually operates above the signal frequency, ie, $f_{LO} = f_{RF} + f_{IF}$. So, for example, a superhet receiver with a 1.6MHz IF tuned to receive a signal at 5.5MHz will operate with an LO at (5.5MHz + 1.6MHz) = 7.1MHz.

As another example, a VHF Band II FM receiver with a 10.7MHz IF covering the broadcast band of 88MHz to 108MHz would require a local oscillator tuning over a range extending from 98.7MHz to 118.7MHz.

As well as the IF signal, many other frequency components will be present at the output of the mixer stage. These will need to be removed so that only the required IF signal is passed on to the IF amplifier stage. This can be achieved using an appropriately designed filter (usually a bandpass type) to achieve the desired selectivity. This topic will be explored in a future Hands-On project.

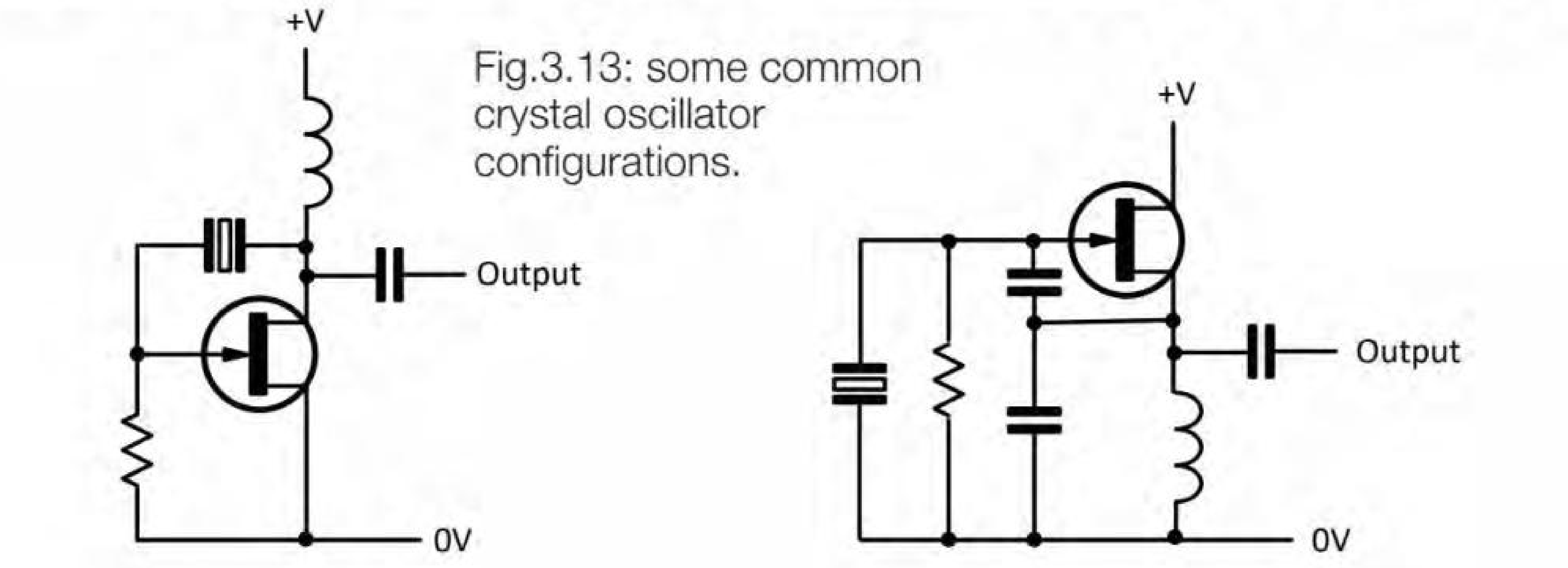
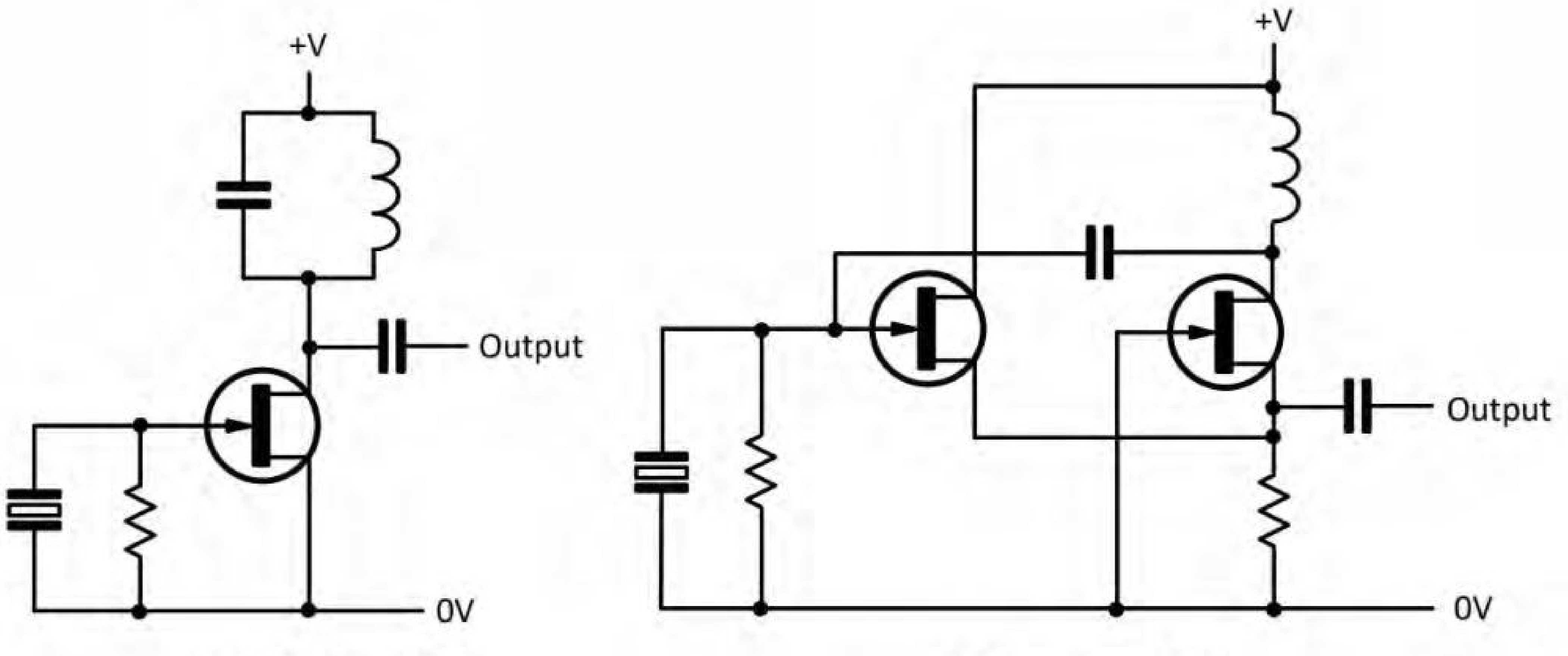


Fig.3.13: some common crystal oscillator configurations.

(a) Pierce oscillator

(b) Colpitts oscillator



(c) Miller oscillator

(d) Franklin oscillator

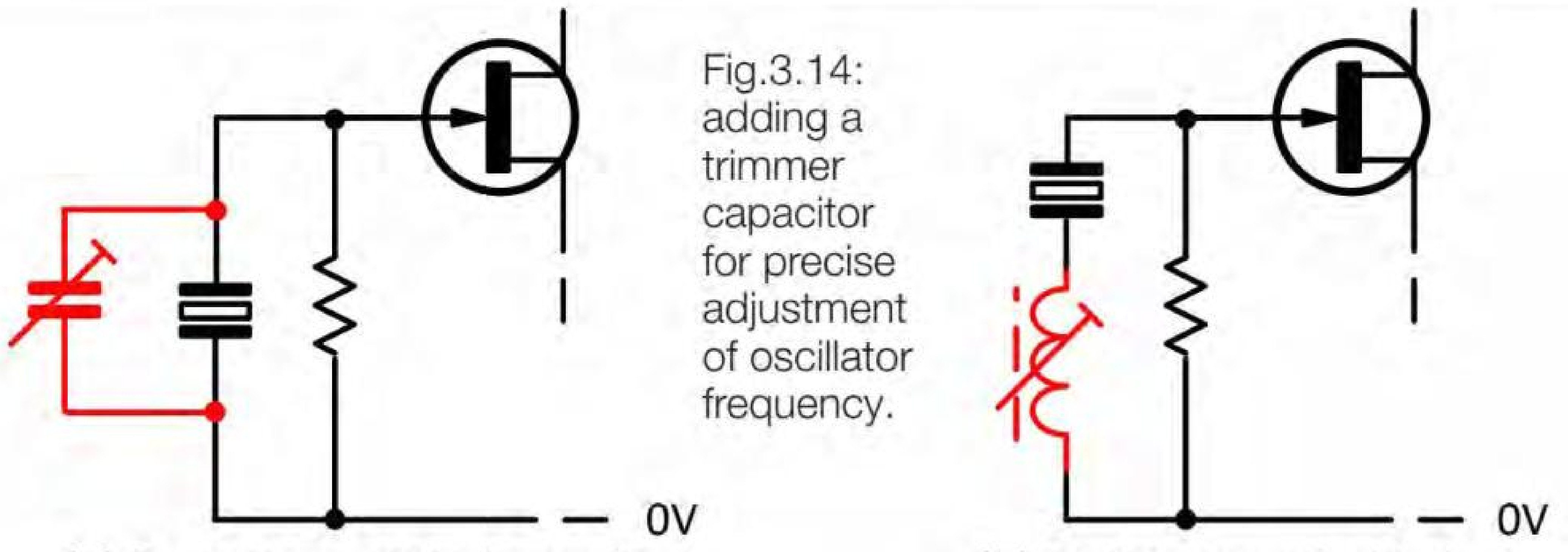


Fig.3.14: adding a trimmer capacitor for precise adjustment of oscillator frequency.

(a) Parallel connected capacitor

(b) Series connected inductor

### Hands-On: An accurate reference oscillator

The first of this month's Hands-On projects features a useful quartz crystal-controlled reference oscillator. This module will provide you with a highly stable frequency source that can be used with several of our future Hands-On projects.

Four basic crystal oscillator configurations are shown in Fig.3.13. For simplicity, a junction gate field-effect transistor (JFET) has been used as the active device. Where a bipolar junction transistor (BJT) is employed, additional bias components will be needed.

The Pierce oscillator in Fig.3.13(a) is a fundamental-mode oscillator with feedback between the drain and gate and the internal capacitance of the device.

The Colpitts circuit shown in Fig.3.13(b) is another popular and reliable fundamental-mode oscillator, and the inductor in the source is often replaced by a resistor.

The Miller oscillator in Fig.3.13(c) is suitable for fundamental as well as overtone operation, depending on the resonant frequency of the L-C drain load. To achieve a precise output frequency, there's often a need to trim the frequency of oscillation. This is usually achieved by a small-value series preset capacitor connected in parallel or a preset inductor connected in series with the crystal (see Fig.3.14).

The circuit of our Hands-On reference oscillator module is shown in Fig.3.15. We have chosen the Pierce configuration, but added a second stage that acts as a unity-gain buffer to provide a relatively high degree of isolation of the load. This helps maintain amplitude and frequency stability when the load impedance changes.

Enhancement-mode insulated-gate FET TR1 (a Mosfet) has a 10MHz quartz crystal connected between its gate and drain in the Pierce configuration, similar to Fig.3.13(a). Trimmer capacitor TC1 provides adjustment of the frequency, while TR2, another 2N7000 Mosfet, acts as a source follower.

DC bias derived from the source of TR2 is fed back to the gate of TR1 via R1. A simple shunt zener stabiliser, D1, helps maintain a constant 6.8V supply to TR1 and TR2. The output amplitude is adjusted via a preset potentiometer, RV1.

The component layout for the reference oscillator module (viewed from the top) is shown in Fig.3.16, while the corresponding track layout (viewed from below) is in Fig.3.17. The required track breaks can be made using a spot face cutter or small drill bit, and the links on the upper side of the boards are made using short lengths of tinned copper wire.

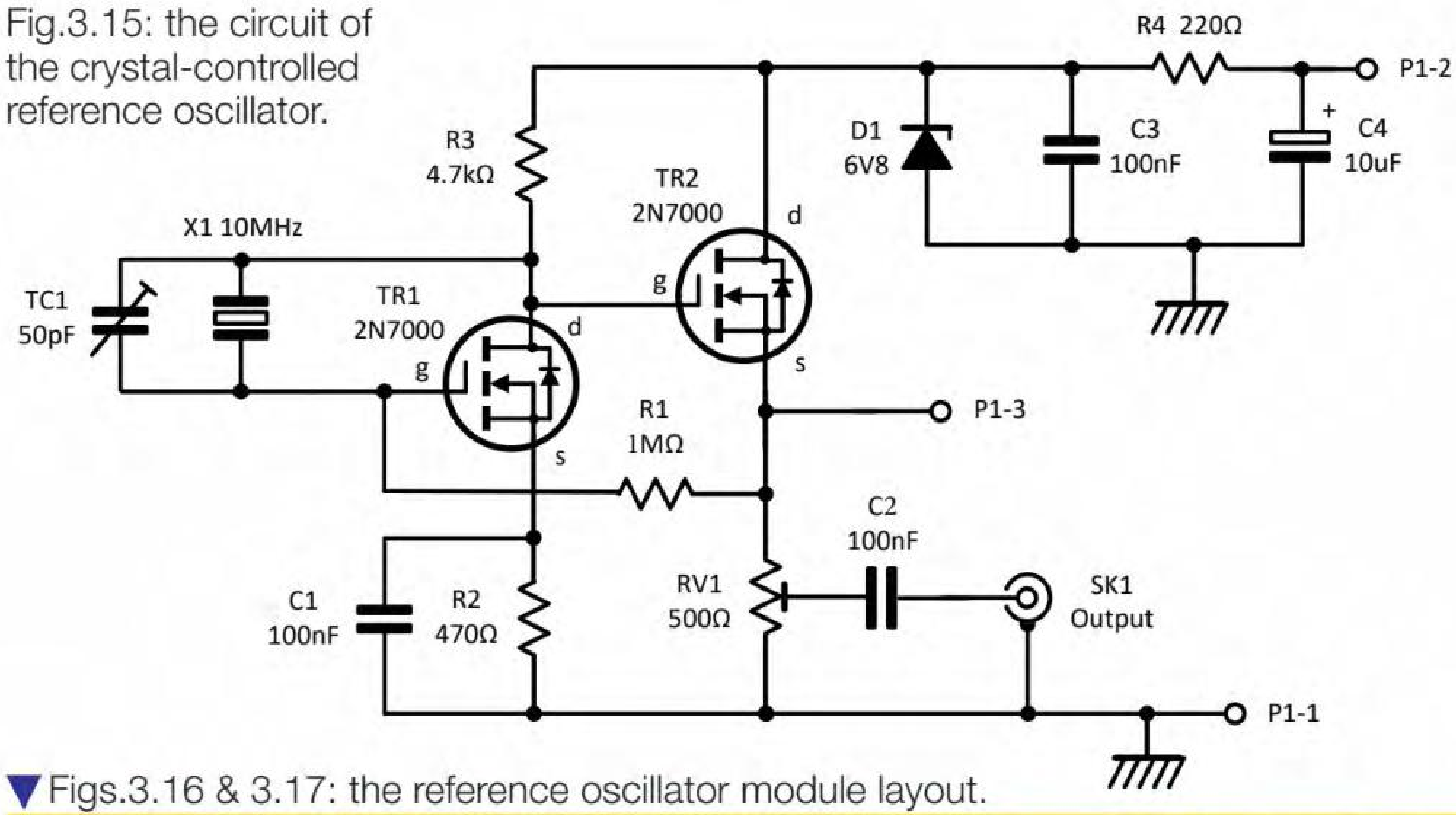
The pin connections for the semiconductor devices are shown later in Fig.3.22. When completed, the stripboard module, DC jack connector and BNC output connector (with pigtail cable and male SMA connector) can be mounted into an ABS enclosure of your choice, as shown in Fig.3.18.

**Testing**

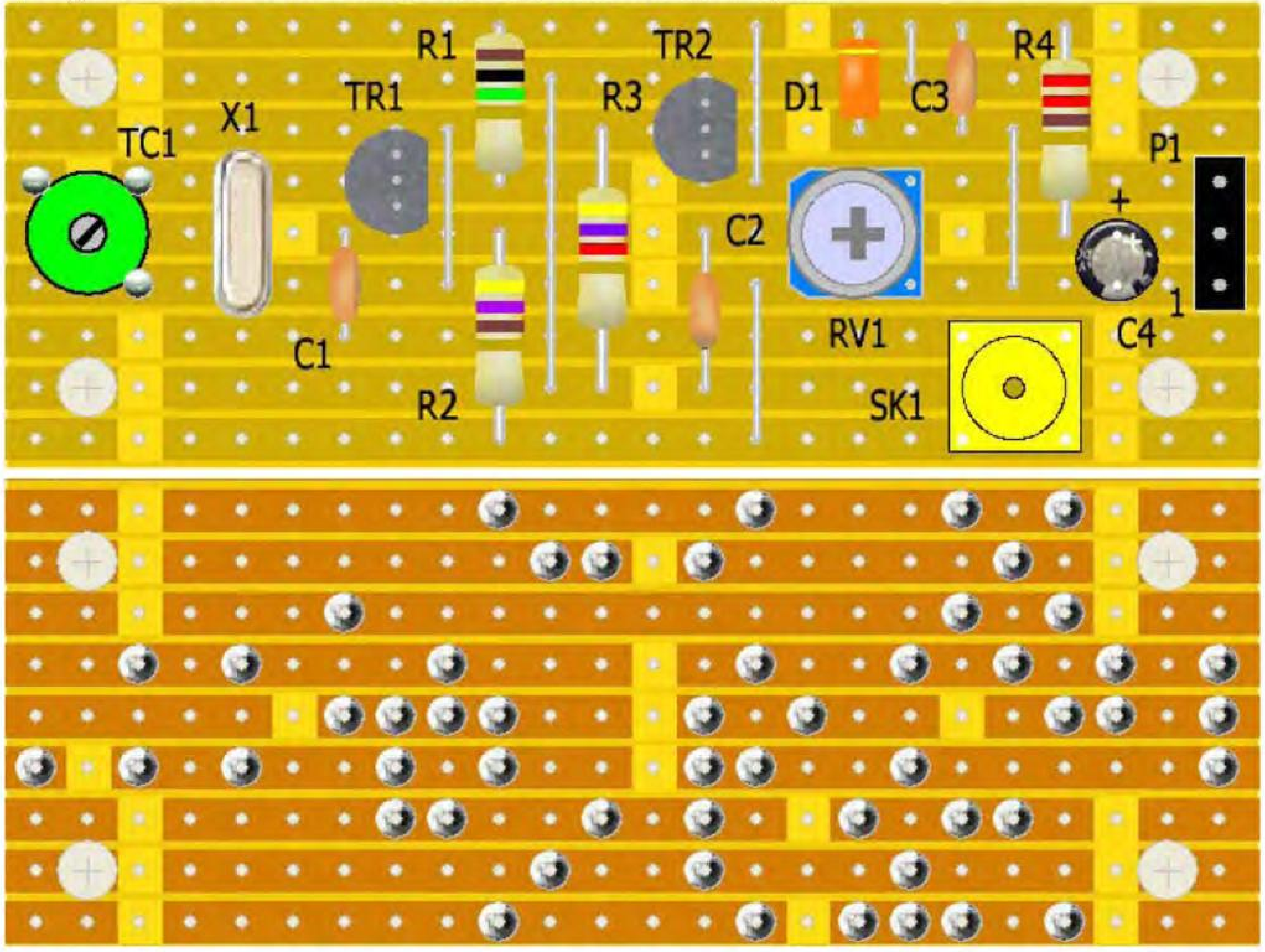
As always, it's important to check the stripboard and internal wiring before applying power. When these checks are complete, apply power to the module and connect the BNC output to an oscilloscope or digital frequency meter. Check the output waveform for frequency (10MHz within a few Hz) and amplitude (about 2V peak-to-peak). Adjust TC1 for an output of exactly 10MHz.

If you don't have any RF test equipment, you can check the oscillator using an ordinary FM receiver. Connect a short length of wire to the BNC output (centre conductor only) and position the receiver close to the oscillator. Now tune the receiver to around 90MHz (or 100MHz).

Fig.3.15: the circuit of the crystal-controlled reference oscillator.



▼Figs.3.16 & 3.17: the reference oscillator module layout.



## Parts List – Reference oscillator module

1 25 × 64mm piece of stripboard (9 × 24 holes)
1 3-pin male 0.1in/2.54mm header (P1)
1 female SMA PCB-mounting connector (SK1)
1 10MHz HC-49S quartz crystal (X1)

**Semiconductors**
2 2N7000 Mosfets, TO-92 (TR1, TR2)
1 6.8V 300-500mW zener diode, DO-35 (D1)

**Capacitors**
1 10µF 16V radial electrolytic (C4)
3 100nF 50V ceramic (C1-C3)
1 50pF miniature preset trimmer (TC1)

**Resistors (all ¼W axial, ±5% or better)**
1 1MΩ (R1)
1 470Ω (R2)
1 4.7kΩ (R3)
1 220Ω (R4)
1 500Ω miniature preset potentiometer/trimpot (RV1)

**Off-board components**
1 ABS enclosure (see text)
1 15cm bulkhead panel-mount BNC female to SMA male coaxial cable adaptor
1 panel-mount 5.5mm DC barrel jack connector
4 M3-tapped, 10mm-long brass or nylon hex spacers
8 M3 × 6mm panhead machined screws

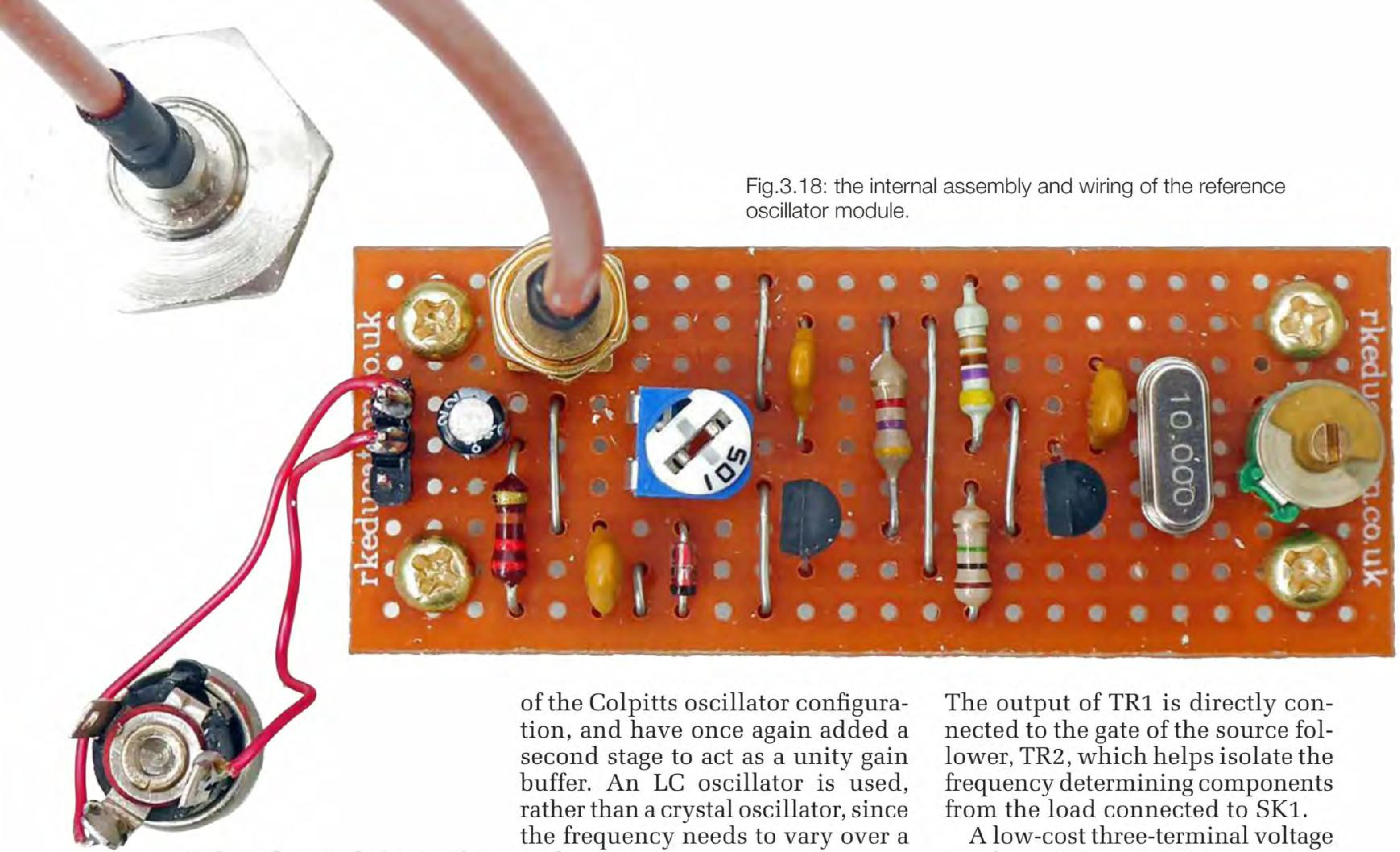

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Fig.3.18: the internal assembly and wiring of the reference oscillator module.

When the ninth (or tenth) harmonic is located, the receiver will go quiet as the strong unmodulated signal is received.

If no output is detected, disconnect the power and recheck the module and off-board wiring. Table 3.1 shows the test voltages obtained from our prototype.

### Hands-On: A variable frequency oscillator (VFO)

Variable frequency oscillators (VFO) are found in RF equipment in applications where continuous variable tuning is required. The circuit of our Hands-On VFO is shown in Fig.3.19. Here, we are making use of the Colpitts oscillator configuration, and have once again added a second stage to act as a unity gain buffer. An LC oscillator is used, rather than a crystal oscillator, since the frequency needs to vary over a wide range.

An NPN silicon bipolar junction transistor, TR1, is connected in Colpitts configuration with its operating frequency determined by L1 and VC1, similar to Fig.3.13(b). The output of TR1 is directly connected to the gate of the source follower, TR2, which helps isolate the frequency determining components from the load connected to SK1.

A low-cost three-terminal voltage regulator, IC1, maintains a constant 9V supply to TR1 and TR2. As before, the output amplitude is adjusted via a preset potentiometer, RV1.

The component layout for the variable frequency oscillator module

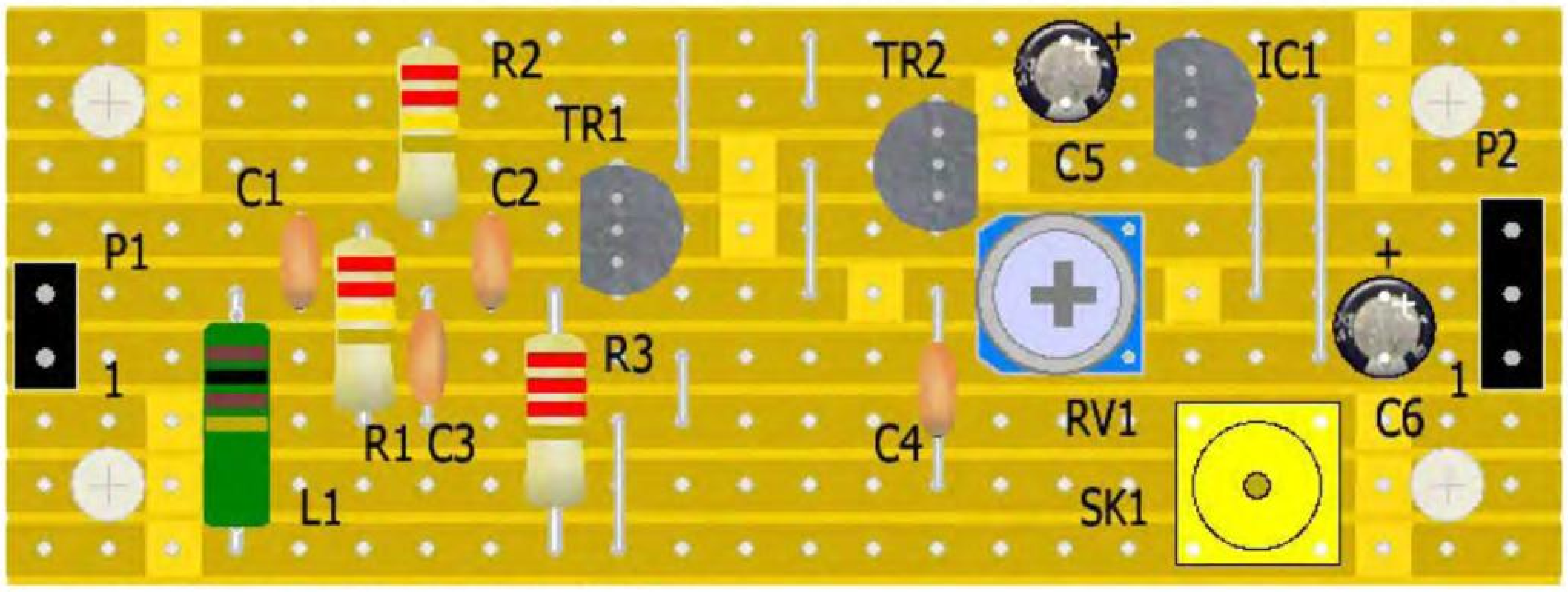

R2
TR1
C1
C2
P1
1
R1 C3
R3
L1
TR2
C5
IC1
P2
C4
RV1
SK1
C6
1
+
+

Fig.3.20: the component layout for the variable frequency oscillator module.

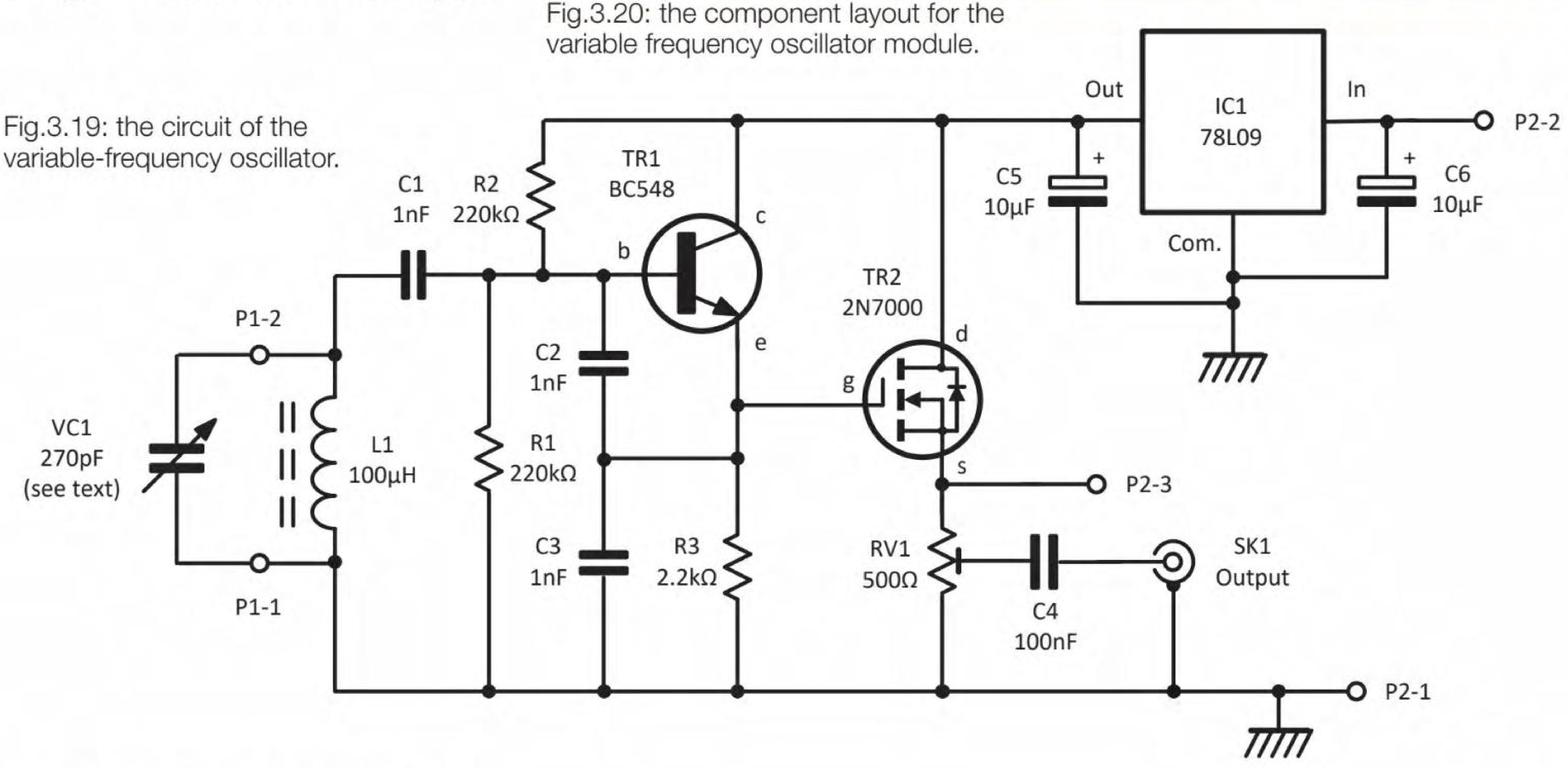

C1
1nF
R2
220kΩ
TR1
BC548
b
c
e
P1-2
VC1
270pF
(see text)
L1
100µH
R1
220kΩ
C2
1nF
C3
1nF
R3
2.2kΩ
P1-1
TR2
2N7000
d
g
s
P2-3
RV1
500Ω
C4
100nF
SK1
Output
Out
IC1
78L09
In
Com.
C5
10µF
C6
10µF
+
+
P2-2
P2-1

Fig.3.19: the circuit of the variable-frequency oscillator.

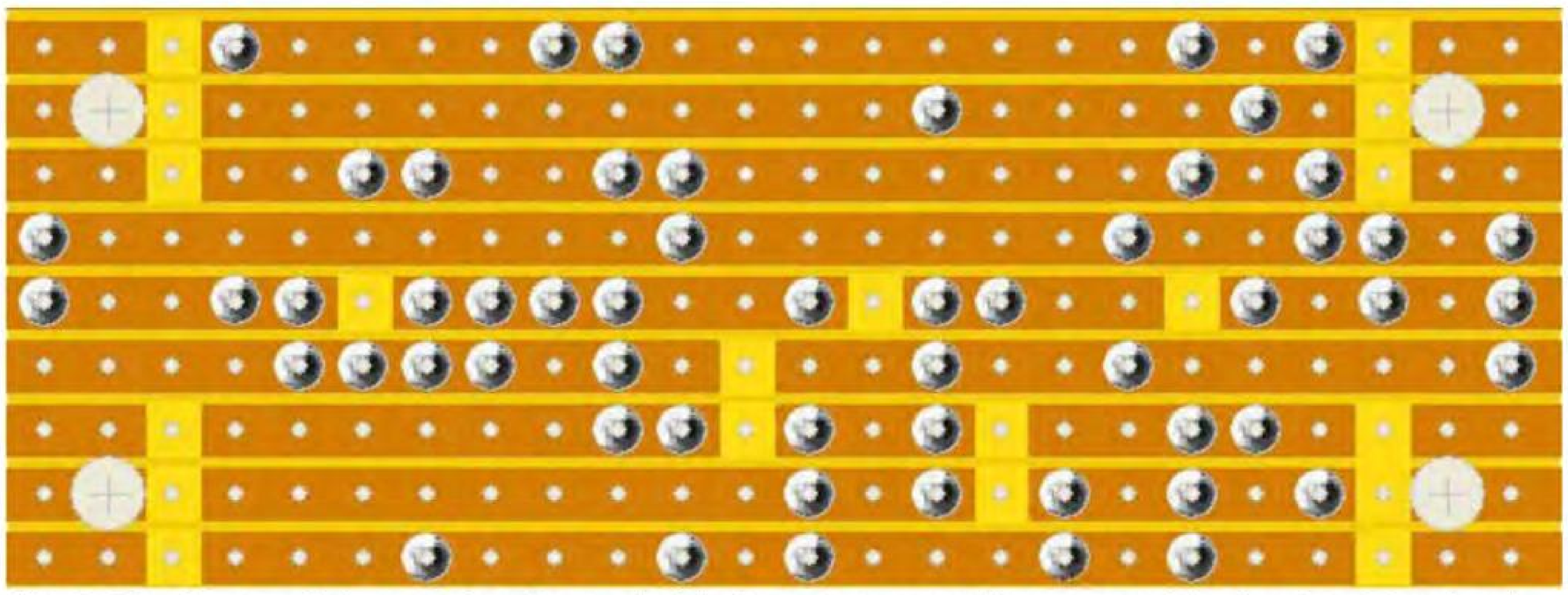
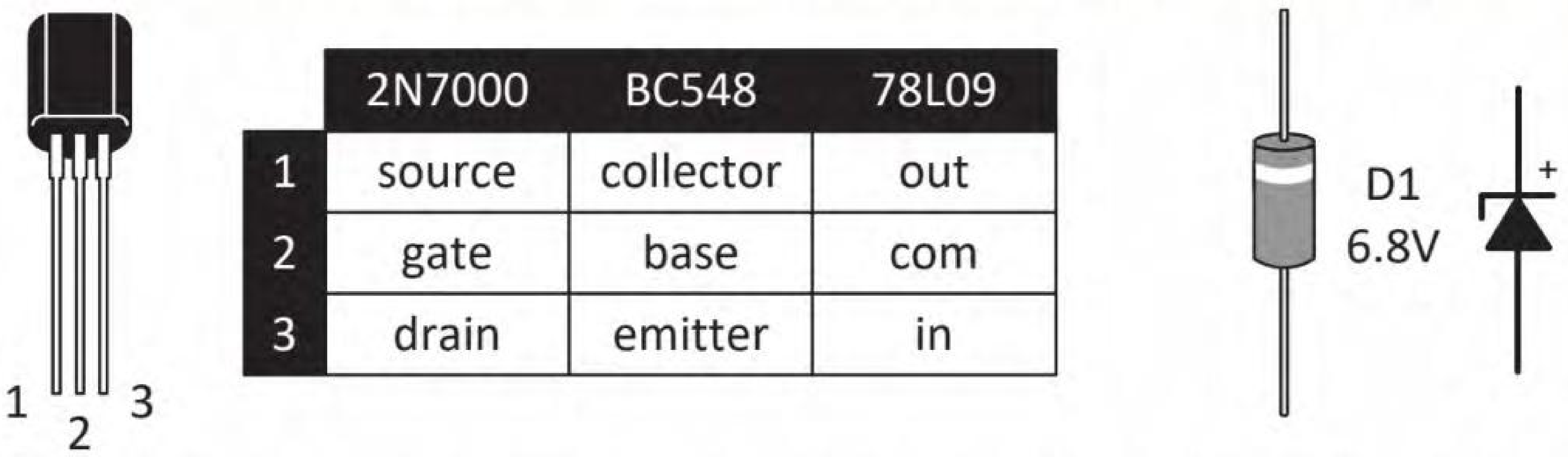


Fig.3.21: the track layout for the variable frequency oscillator module (underside view).



| | 2N7000 | BC548 | 78L09 |
|---|---|---|---|
| 1 | source | collector | out |
| 2 | gate | base | com |
| 3 | drain | emitter | in |

Fig.3.22: the pin connections for the semiconductors used in both Hands-On projects.

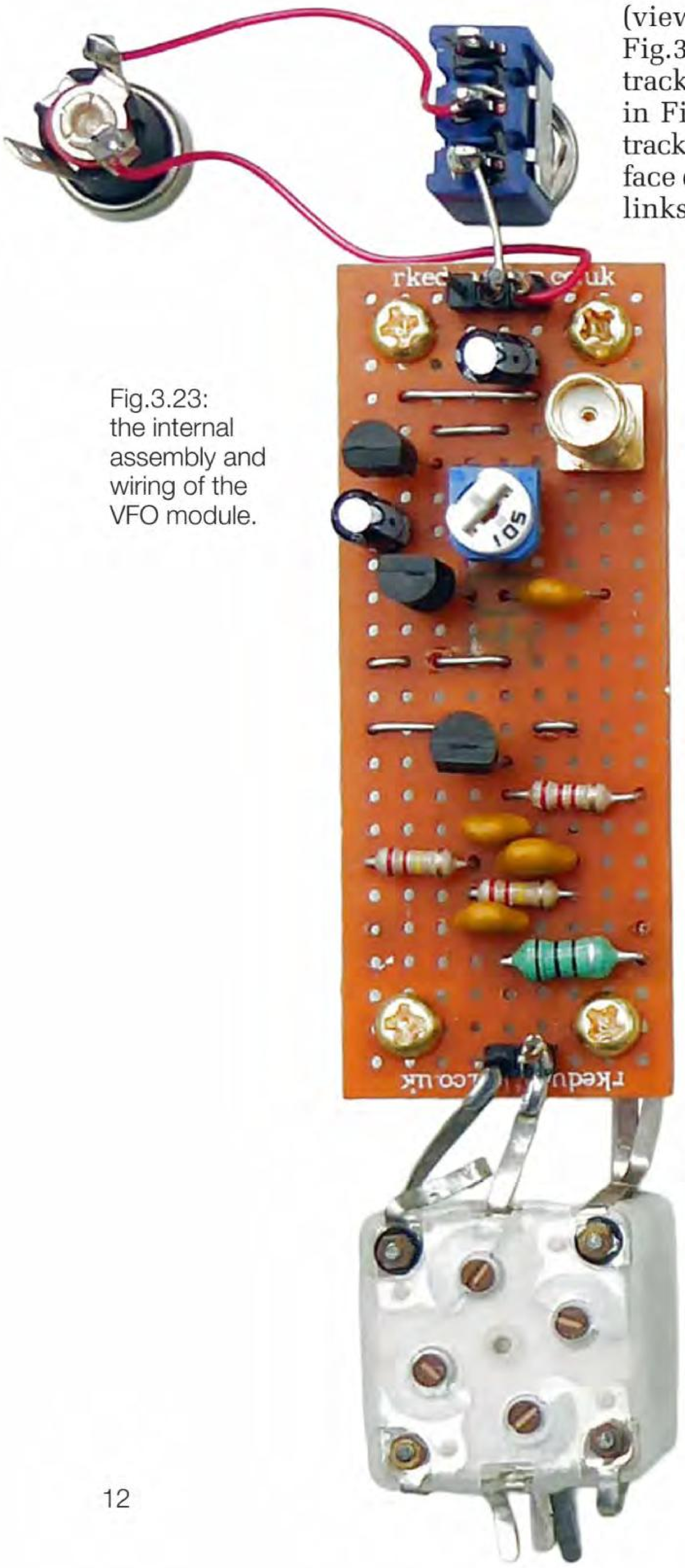


Fig.3.23: the internal assembly and wiring of the VFO module.

(viewed from the top) is shown in Fig.3.20, while the corresponding track layout (viewed from below) is in Fig.3.21. As usual, the required track breaks can be made using a spot face cutter or small drill bit, and the links on the upper side of the boards are made using short lengths of tinned copper wire.

The pin connections for the semiconductor devices are shown in Fig.3.22. When completed, the stripboard module, on/off switch, DC jack connector and BNC output connector (with pigtail cable and male SMA connector) can be mounted into another small ABS enclosure of your choice (Fig.3.23) or retained for use with next month's Hands-On project.

**Testing**

Once again, it's important to check the stripboard and internal wiring before applying power. When these checks are complete, apply 12V to the DC jack, switch it on, set RV1 to mid-position and connect the BNC output to an oscilloscope or digital frequency meter.

Observe the output waveform as VC1 is varied over its full range. The frequency should be adjustable over a range extending from about 680kHz (with VC1 full anticlockwise) to 910kHz (with VC1 fully clockwise). Note that the VFO tuning range will depend on the component used for VC1 (smaller values will produce a correspondingly smaller VFO tuning range).

**Table 3.1 – Expected voltages (Ref. Osc.)**

| Device | Pin | Voltage |
|---|---|---|
| TR1 | D | 2.8V |
| | G | 1.3V |
| | S | 0.5V |
| TR2 | D | 6.8V |
| | G | 2.8V |
| | S | 0.8V |

**Table 3.2 – Expected voltages (VFO)**

| Device | Pin | Voltage |
|---|---|---|
| TR1 | C | 9.1V |
| | B | 2.8V |
| | E | 4.9V |
| TR2 | D | 9.1V |
| | G | 4.9V |
| | S | 3.1V |
| IC1 | IN | 12V |
| | COMMON | 0V |
| | OUT | 9.1V |

Finally, with RV1 set to maximum (fully clockwise), check that the output is about 2V peak-to-peak over the full VFO tuning range.

If you don't have any RF test equipment, the VFO can be tested with the aid of an ordinary medium-wave AM receiver (including either of the AM receivers that we described in previous parts of this Teach-In series).

Connect a short length of wire to the BNC output (centre conductor only) and place the receiver within about two metres (about six feet) of the VFO. With VC1 set to its mid position, tune the receiver between about 700kHz and 900kHz. When the VFO signal is located, the receiver will go quiet as the strong unmodulated signal is received.

Next, locate a strong MW broadcast station in the same range (in the UK you can use BBC Radio 5 Live on 693kHz and 909kHz for this test). Having located the broadcast signal, rotate VC1 until you hear a loud whistle (beat note) as the VFO is tuned across the broadcast signal. This neatly illustrates the BFO effect that we discussed earlier!

If no output is detected, switch it off and disconnect the power before rechecking the module and off-board wiring and, in particular, the connections to VC1. If you are still experiencing problems, Table 3.2 provides the test voltages obtained from our prototype.

## Parts List – Variable frequency oscillator

1 25 × 64mm piece of stripboard (9 × 24 holes)
1 2-pin male 0.1in/2.54mm header (P1)    1 3-pin male 0.1in/2.54mm header (P2)
1 female SMA PCB-mounting connector (SK1)
1 100µH axial RF inductor (L1)

**Semiconductors**
1 BC548 30V 100mA NPN bipolar junction transistor, TO-92 (TR1)
1 2N7000 Mosfet, TO-92 (TR2)
1 78L09 9V linear voltage regulator, TO-92 (IC1)

**Capacitors**
2 10µF 16V radial electrolytic (C5, C6)
1 100nF 50V ceramic (C4)
3 1nF 50V ceramic (C1, C2, C3)
1 270pF miniature solid-dielectric variable (VC1)

**Resistors (all ¼W axial, ±5% or better)**
2 220kΩ (R1, R2)    1 2.2kΩ (R3)
1 500Ω miniature preset potentiometer/trimpot (RV1)

**Off-board components**
1 ABS enclosure (see text)
1 SPST miniature toggle switch
1 15cm bulkhead panel-mount BNC female to SMA male coaxial cable adaptor
1 panel-mount 5.5mm DC barrel jack connector
4 M3-tapped, 10mm-long brass or nylon hex spacers
8 M3 × 6mm panhead machined screws

### Using software simulation

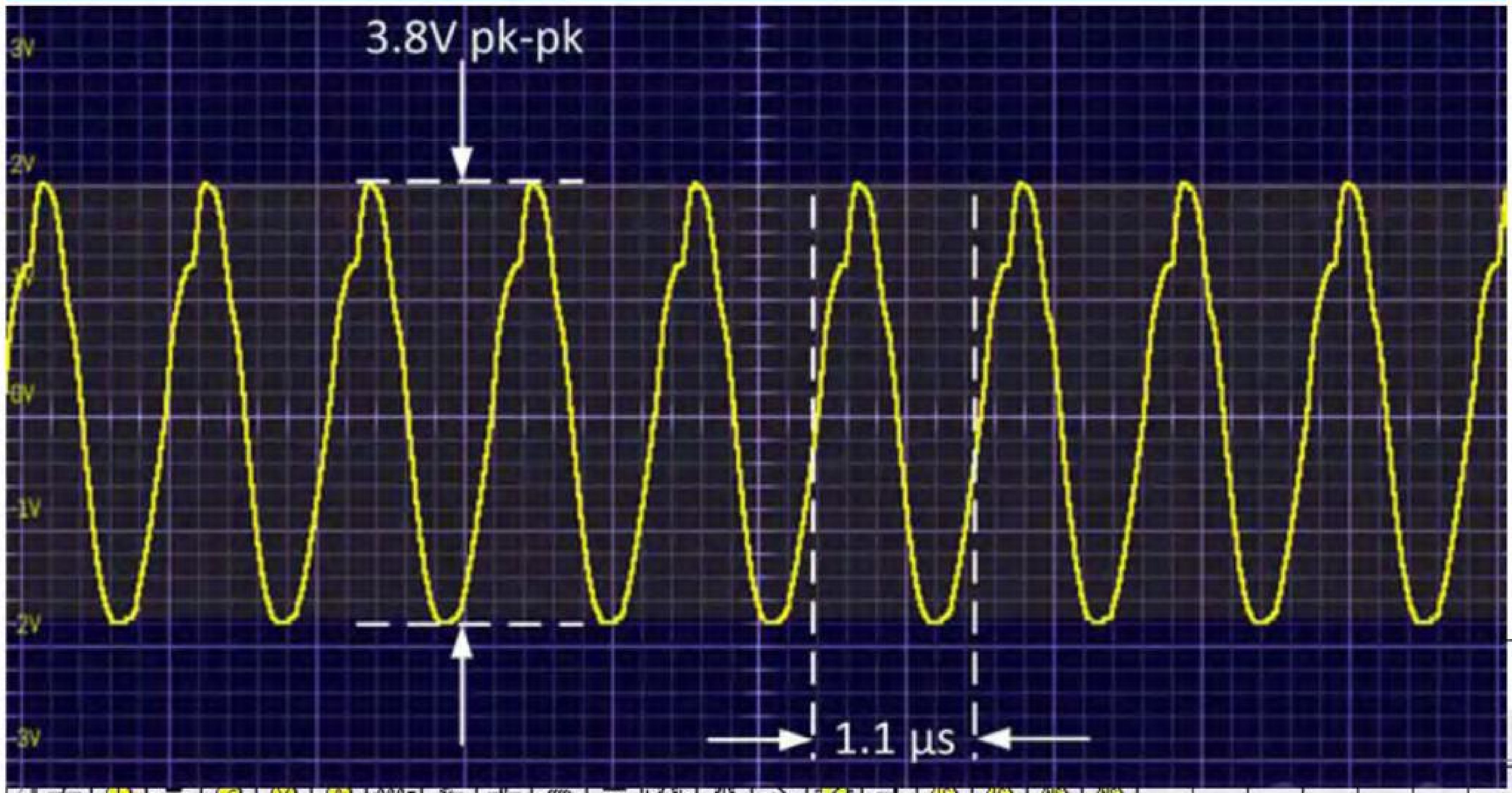
Simulation software can be invaluable in most areas of RF design. As an example, we used software modelling to test our VFO module, comparing the simulated results with those obtained from real measurements. Fig.3.24 shows the module simulated using the popular (and freely available) Tina-TI SPICE package from Texas Instruments (available from https://www.ti.com/tool/TINA-TI).

It's worth comparing Tina's virtual oscilloscope display in Fig.3.24 with that obtained using a real digital storage scope (Fig.3.25). The two waveform displays are virtually identical, illustrating the value of software modelling techniques for this simple RF application.

Component values and active devices can be very easily changed, and the results observed using Tina's handy range of virtual instruments.

### Output spectrum and harmonics

If you take a close look at Fig.3.25, you should notice that the output of our VFO is not perfectly sinusoidal. This distortion results from nonlinearity, and it indicates the presence of harmonic components in the output from the VFO. These can be viewed and analysed using Fast Fourier Transform (FFT) techniques provided by the DSO software.


3.8V pk-pk
3V
2V
0V
-1V
-2V
-3V
1.1 µs

◀ Fig.3.25: the output waveform from the VFO module displayed on a low-cost digital storage oscilloscope (DSO).

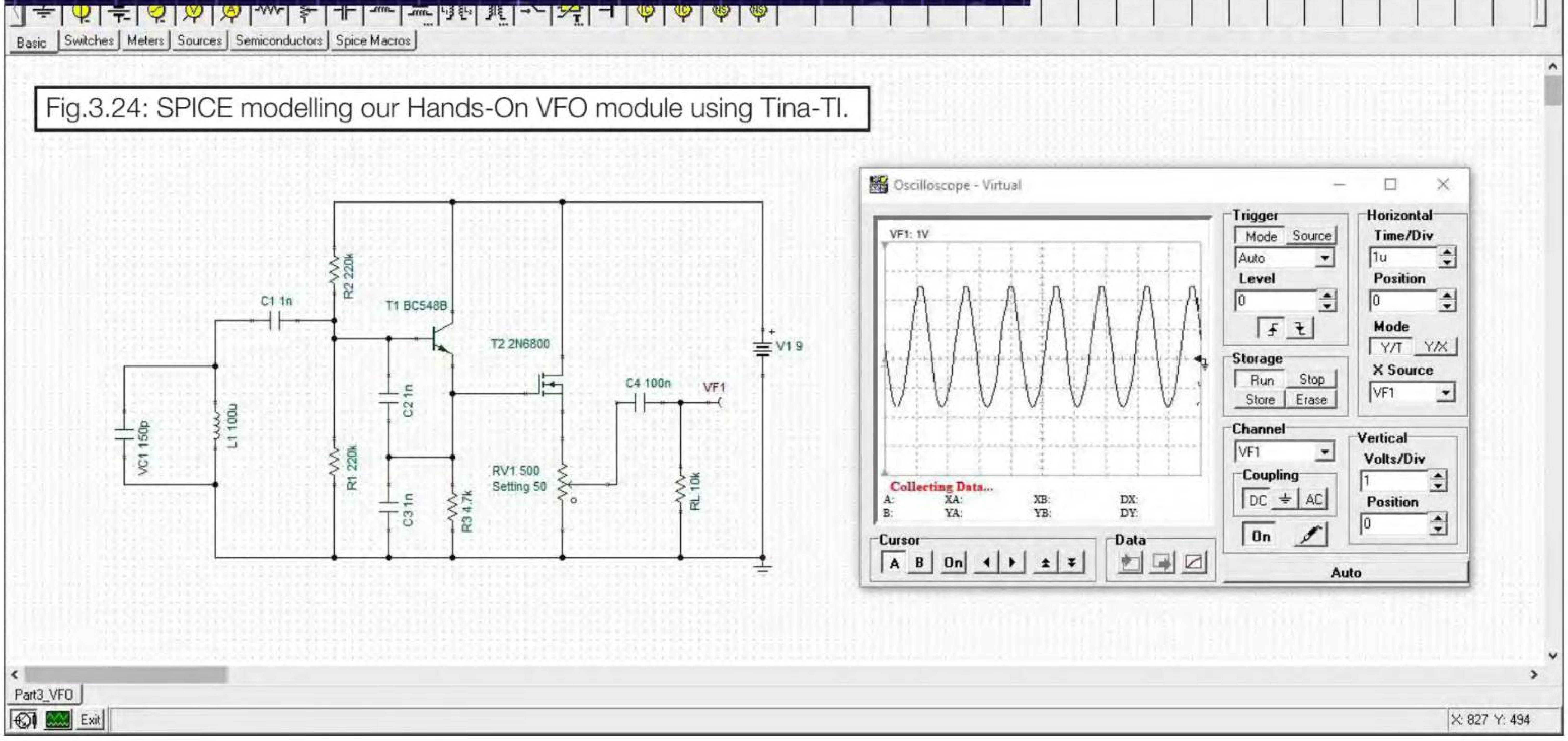


Fig.3.24: SPICE modelling our Hands-On VFO module using Tina-TI.
Ground
Basic
Switches
Meters
Sources
Semiconductors
Spice Macros
C1 1n
R2 220k
T1 BC548B
T2 2N6800
V1 9
C2 1n
C4 100n
VF1
VC1 150p
L1 100u
R1 220k
C3 1n
R3 4.7k
RV1 500
Setting 50
RL 10k
Oscilloscope - Virtual
VF1: 1V
Collecting Data...
Trigger
Mode
Source
Auto
Level
0
Storage
Run
Stop
Store
Erase
Channel
VF1
Coupling
DC
AC
On
Horizontal
Time/Div
1u
Position
0
Mode
Y/T
Y/X
X Source
VF1
Vertical
Volts/Div
1
Position
0
Cursor
A
B
On
Data
Auto
Part3_VFO
Exit
X: 827 Y: 494

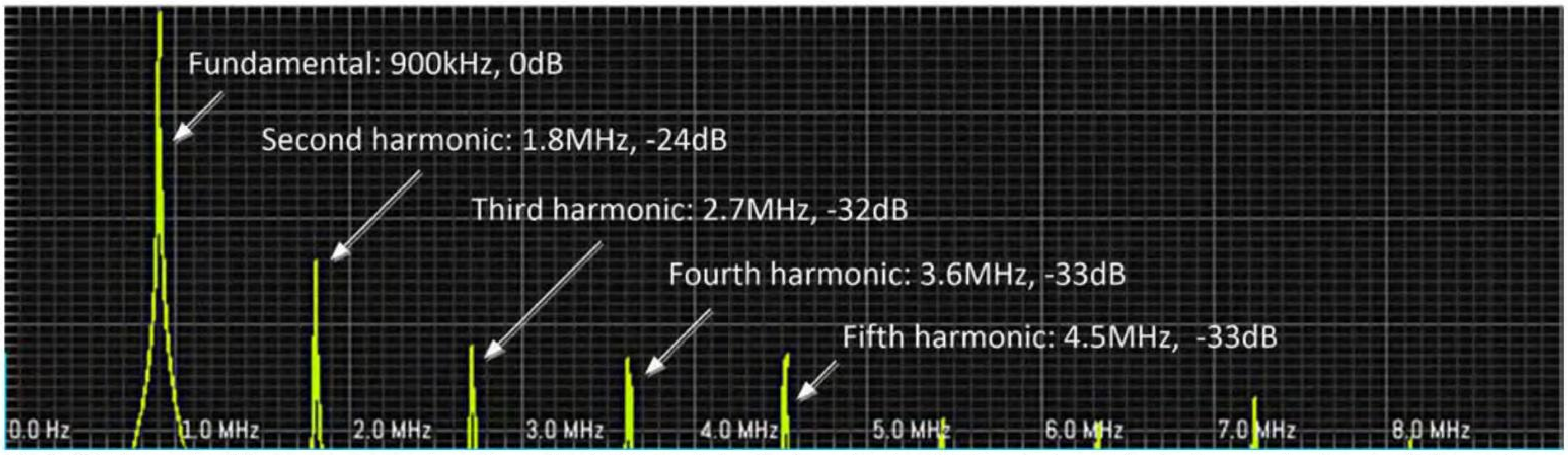


Fig.3.26: the output frequency spectrum from the VFO module displayed using FFT techniques (note the relative levels of the various harmonic components).

The FFT-derived spectrum of the VFO output is shown in Fig.3.26. We've indicated the order of each of the most significant harmonics, as well as their relative level (shown in decibels, dB) below the fundamental. Notice how, in general, the amplitude of the harmonic components decreases with the order of the harmonic.

Note how the second harmonic is at double the frequency of the fundamental, the third harmonic is at triple the frequency and so on. You can see harmonics up to the 9th here; in general, they decay in amplitude as they increase in frequency, although not always.

It's important to note that harmonics present in the output of a VFO (or any other oscillator for that matter) are generally undesirable. In a receiver, they can cause unwanted mixing products. In a transmitter, they can result in the emission of signals that may cause interference to other services.

In practice, we would either need to improve the oscillator design or to apply subsequent filtering to ensure that any harmonic components are reduced to a negligible level.

**Coming up!**

In Part 4 of this Teach-In series, we will be delving into the exciting world of software-defined radio (SDR).

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Our Hands-On project next month will put this month's RF oscillator modules to good use as the basis for a complete amplitude-modulated signal source. **PE**

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# High-Bandwidth Differential Probe

This high-bandwidth, high-voltage differential probe is ideal for use with oscilloscopes, although it could have other uses. It has an internal rechargeable battery and fits in the same case as the Isolated Current Probe we published in the November issue. It will be an invaluable addition to your test equipment arsenal!

**By Andrew Levido**

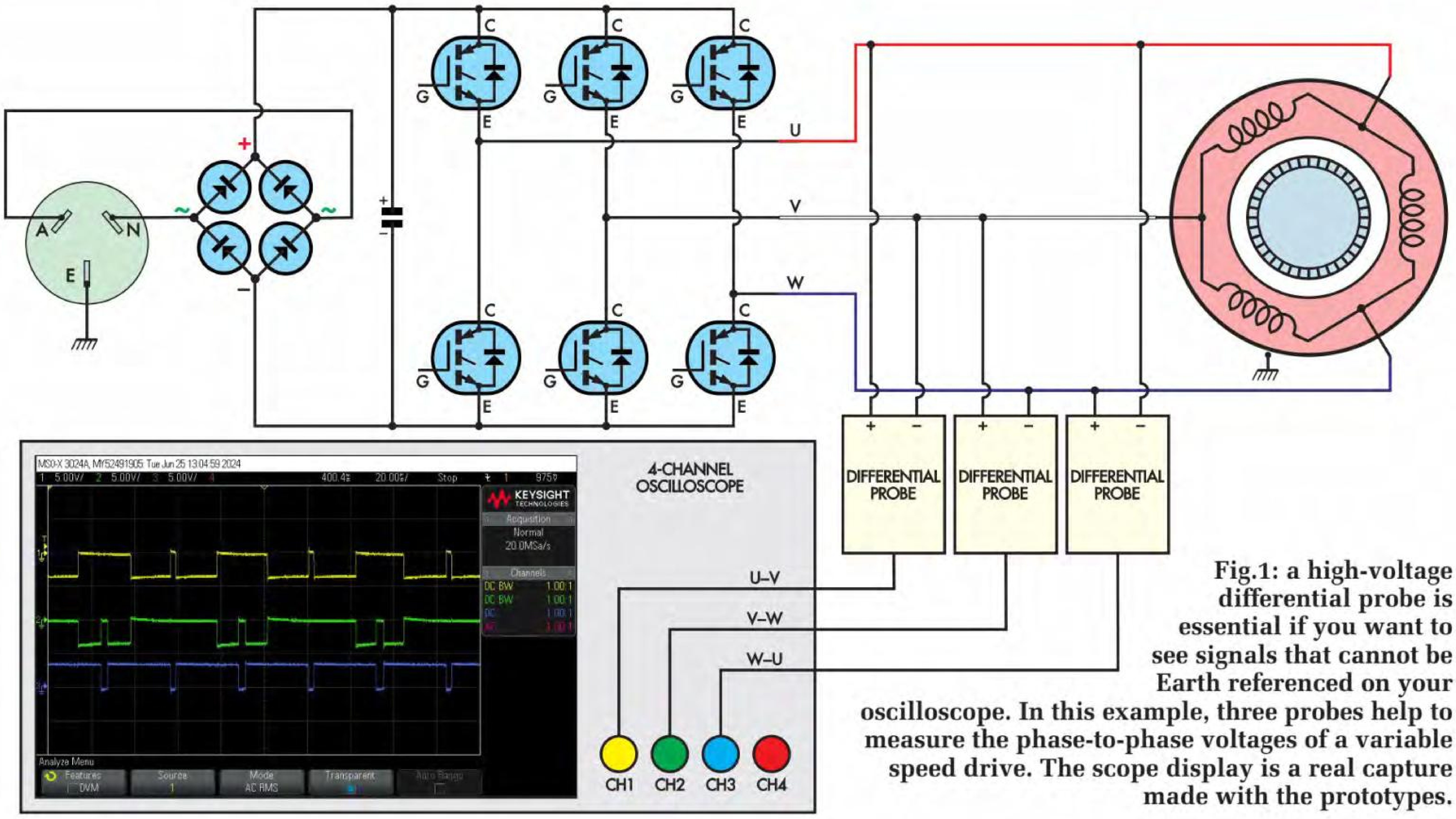
**If you ever work with high-voltage circuits,** a differential probe is an indispensable piece of test equipment. In fact, they're also useful with many low-voltage circuits; any time you want to monitor a differential voltage between two points in a circuit. This one can be built for a fraction the cost of a commercial device with similar performance and functions.

The ground sides of most oscilloscope inputs are connected directly to mains Earth. This means you can only measure Earth-referenced signals – either those already referenced to Earth, or those that you can safely connect to Earth on one side for the purposes of the measurement. That generally includes truly floating circuits, such as battery-powered devices.

Unfortunately, many signals in circuits such as switch-mode power supplies or motor controllers are referenced to voltages well above Earth potential. Connecting a scope to these using a standard probe would create a short from the circuit reference to mains Earth, via the probe ground lead and the 'scope itself. This will potentially be catastrophic for your scope, the probe and your circuit.

Even if your circuit is floating and you can safely Earth one point for testing, if you want to measure another voltage at the same time that's referenced to a different point, you're out of luck. That's because if you Earth two different points in your circuit, you are adding a short circuit; usually not a great idea! A differential probe (or multiple probes) totally solves that problem.

As an interesting and slightly terrifying aside, my very first oscilloscope, an Australian made BWD830 purchased in the early 1980s, actually has a "ground isolate" switch on the back panel that allows the user to open the mains Earth connection, allowing the scope common to float.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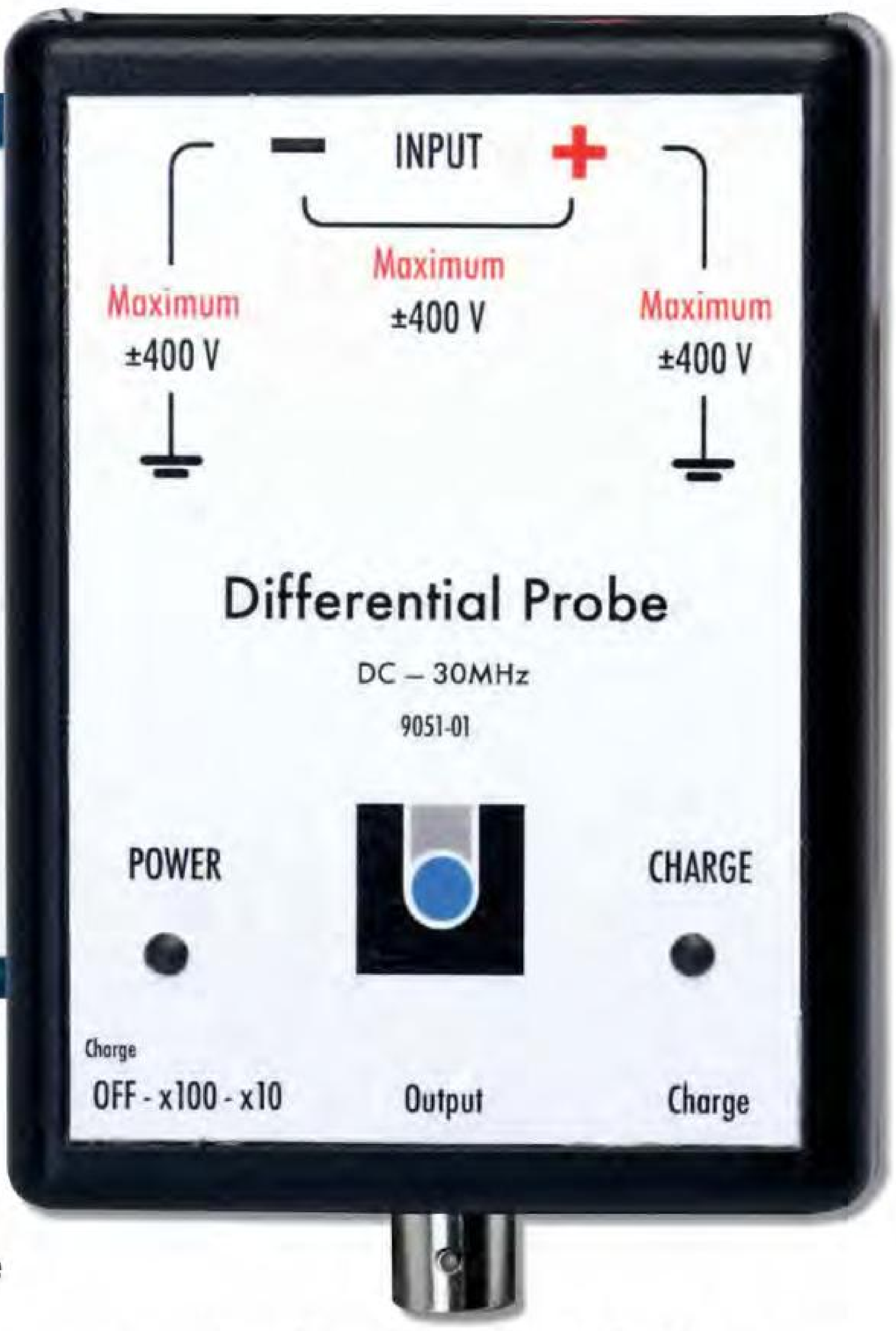
DC 1.00:1
AC 1.00:1
Analyze Menu
Features DVM
Source 1
Mode AC RMS
Transparent
Auto Range

**Fig.1: a high-voltage differential probe is essential if you want to see signals that cannot be Earth referenced on your oscilloscope. In this example, three probes help to measure the phase-to-phase voltages of a variable speed drive. The scope display is a real capture made with the prototypes.**

# High-Bandwidth Differential Probe

- **Maximum common-mode voltage:** ±400V DC (280V RMS)
- **Maximum differential-mode voltage:** ±400V DC (280V RMS)
- **Common-mode input impedance:** 2MΩ || 2.5pF
- **Differential-mode input impedance:** 4MΩ || 2.5pF
- **Attenuation ranges:** 100:1, 10:1
- **Basic DC accuracy:** better than 1%
- **Bandwidth:** >30MHz (x100), >25MHz (x10)
- **CMRR:** >100dB (DC-100Hz)
- **Battery Life:** >4 hours
- **Charging time:** <3 hours
- **Charging socket:** USB-C
- **Input sockets:** 4mm banana sockets, 20mm spacing
- **Output socket:** BNC



This avoids the risk of blowing up the scope, but can allow the scope case and front panel terminals to rise to lethal voltage levels! Thankfully, this dangerous practice is a thing of the past. (And it still doesn't help for monitoring multiple points referenced to different voltages anyway...)

Fig.1 shows an example of where a differential probe is indispensable. Here, the three phase-to-phase PWM output waveforms from a variable speed drive (suspiciously similar to the one we introduced in the December 2025 issue) are displayed on three channels of an oscilloscope.

None of the U, V or W phases can be safely Earthed, and the voltages involved are in the order of 400V peak-to-peak. The differential probes provide 100:1 attenuation of the differential voltage and over 10,000:1 (100dB) attenuation of the common-mode voltage, allowing the phase-to-phase voltages to be measured safely.

The waveforms shown on the scope are from a real screen capture made using three of these devices.

A high-voltage differential probe translates the difference in voltage between its two high-impedance inputs into a voltage that you can safely connect to your oscilloscope's Earthed input. The output is proportional to the difference in voltage between the positive and negative inputs. Any common-mode signal present on both inputs is almost entirely rejected.

The differential probe is housed in a small plastic case measuring 82 × 65 × 28mm. The inputs are two shrouded 4mm banana jacks at one end, with a 20mm spacing. That is close enough to the ¾-inch (19.05mm) standard for dual-banana-plug accessories to fit. The BNC output, range switch and USB-C connector are at the other end of the case. The power and charge LEDs are visible through the top of the case via two light pipes.

## Design goals

When I set out on this project, I set myself a few design goals. I wanted a probe that could safely be used in mains-voltage projects like that described above. This means the device should be able to measure differential-mode signals of ±400V magnitude and withstand a similar level of common-mode voltage. This corresponds to an AC voltage of $280V_{RMS}$.

We want to show these on a standard scope, so an attenuation of 100:1 (-40dB) would be appropriate to give a ±4V full-scale output signal.

Sometimes, we will want to measure a low-voltage signal riding on a high common-mode voltage; for example, to examine the gate signals of the IGBTs in Fig.1. These signals would normally be within a ±40V range, so a 10:1 (-20dB) attenuation range was also one of my requirements.

The common-mode rejection ratio (CMRR) at DC to mains frequency should be at least 100dB. This means a 400V common mode voltage would contribute less than 4mV at the output. The input impedance should be in the megohm range with fairly low parallel capacitance (say <10pF).

I wanted an upper bandwidth limit as high as I could reasonably get, at least 25MHz, to get good representation of high-speed switching signals with fast rise-times. Bandwidth and rise time are related according to the approximation $t_{RISE} \approx 0.35/BW$, so a 25MHz bandwidth means the fastest rise time we will see is about 14ns, which should be short enough.

I also wanted the unit to have the smallest form factor possible and include an internal rechargeable battery. My bench gets cluttered enough as it is without having bulky probes and their power cables added to the mix. More than three hours' battery life and USB-C recharging was mandatory.

## Operating principles

In principle, the concept of a differential probe is pretty straightforward: a matched pair of input attenuators followed by a classic three op-amp differential instrumentation amplifier will do the job. Fig.2 shows the bare bones of the circuit, along with differential and common-mode voltage sources we will discuss later.

You can think of this circuit has having three sections: a dual input attenuator, a buffer stage and a difference amplifier stage. The overall differential-mode gain of the circuit is given by multiplying the gains of each of these stages, which are given in the figure.

We have to set the gains of each stage such that we respect the input common-mode voltage range of each op amp (voltages A+/A- and C+/C- in the figure) and their maximum output swings (voltages B+/B- and $V_{OUT}$). With ±5V power rails, it seemed fairly safe

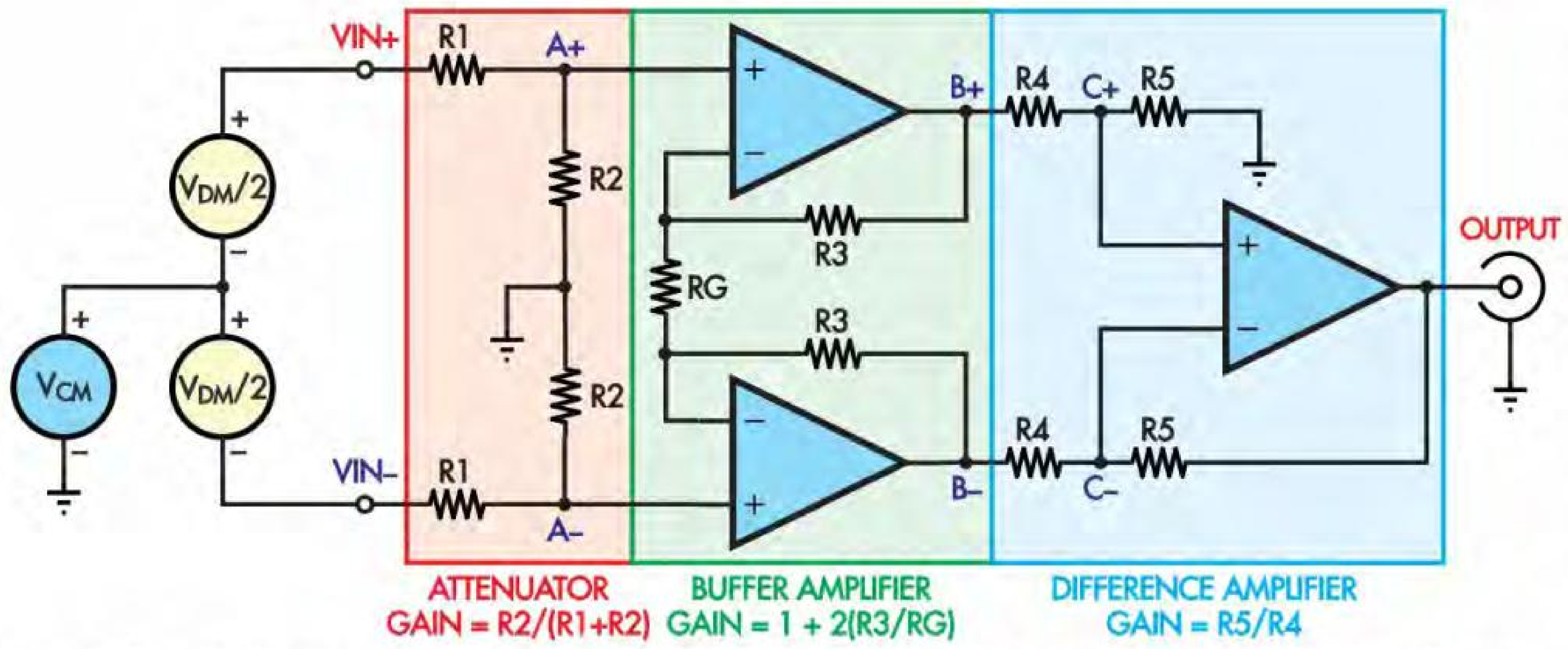


**Fig.2: the differential probe consists of two matched attenuators followed by a classic three-op-amp instrumentation amplifier. The latter has a buffer stage with a gain programmable via a single resistor (RG) and a difference amp stage with a fixed gain.**

to assume an input common-mode voltage of ±2V and an output swing of ±4V as a starting point.

A division ration of 200:1 would give 2V at point A with a 400V input, leaving the rest of the circuit to provide a gain of 2 or 20 to achieve the overall target of 100:1 or 10:1 attenuation. As you can see from Fig.2, the voltage at any one of the inputs will actually be a combination of some common-mode voltage, $V_{CM}$, plus one half of the differential-mode voltage, $V_{DM}$.

The maximum voltage of 400V at the inputs will therefore be made up of a combination of common-mode and differential-mode voltages. We can construct a graph (the yellow area in Fig.3) showing the allowable ranges of input voltage that keep the op amp voltage within the ±2V band.

This input range is more than enough to measure signals likely to be encountered in a circuit powered by 230-240V AC.

The area shown in pink is the combination of inputs that can be measured on the 10:1 attenuation range. In this case, the range is limited by the ±4V output swing of the op amps, rather than the input common-mode voltage.

It is important to keep in mind that all of these are limits relate to the faithful reproduction of the input signal. The maximum voltage that the inputs can safely withstand is considerably higher, as we shall see.

With the attenuator gain determined to be 1/200th, we can consider the gains for the other two stages. The buffer stage gain can be set by selecting a single resistor RG, so this is the obvious candidate for switchable part of the gain. We can't put all the remaining gain in this stage, or we run the risk of exceeding the difference amplifier's common-mode input voltage range.

Thus, I chose to make the buffer stage gain switchable between 1 and 10 and set the difference amp stage gain to a fixed value of two times.

That's about it for the high-level design – a pair of matched 200:1 attenuators, a ×1/×10 switchable buffer stage and a ×2 difference amplifier. Now we just have to make it all work – and the devil is in the details, as they say.

## The attenuator

The circuit diagram (Fig.4) shows the complete design. The attenuators have to withstand high voltages, have reasonably high input impedance and be very closely matched to maximise CMRR. The attenuators are identical, so I will focus this description on the positive side for simplicity.

The resistors I have chosen for the upper leg of this divider are 1MΩ ±0.1% ¼W devices with a voltage rating of 700V. The maximum continuous voltage we can apply across each of these resistors is limited to 500V by power dissipation. With two resistors in series, the inputs can withstand a sustained voltage of 1kV (DC or AC RMS), giving a comfortable safety margin.

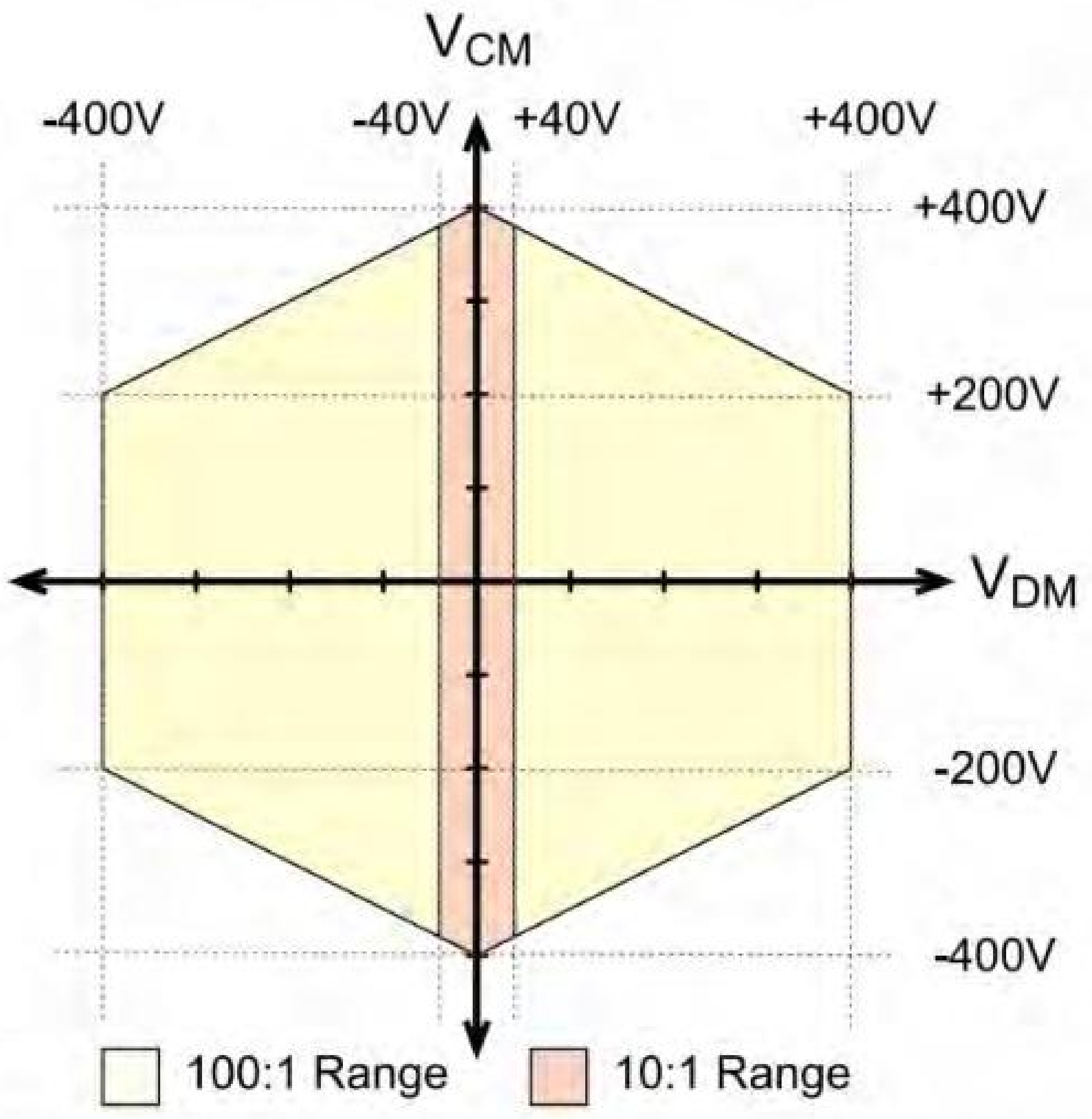


**Fig.3: the range of common-mode voltage and differential-mode voltages the probe can faithfully reproduce at the output. The yellow area is for the ×100 range and the pink area is for the ×10 range. It can safely tolerate much higher voltages without damage.**

The resistance value required in the bottom leg of the divider for 200:1 attenuation is 10,050Ω. In each half, this is made up of a 10kΩ resistor, a 10Ω resistor and half of 100Ω trimpot VR1. This trimpot is shared with the negative attenuator, allowing us to tweak the divisor ratios so that they are precisely equal, as necessary for maximum rejection of common-mode signals.

With VR1 centred, the total resistance of each resistor string is 10,060Ω, not 10,050Ω as calculated. The extra 10Ω is necessary to compensate for the 10MΩ resistors, which are effectively in parallel with the lower leg of each divider. You can ignore trimpot VR2 in this calculation, since its value is much smaller than the error due to the 1% tolerance in the value of the 10MΩ resistors.

The overall resistance of the lower leg of each divider is therefore 10,060Ω || 10MΩ = 10,050Ω.

The purpose of VR2 is to allow us to inject up to ±5mV into one input to compensate for any op amp offset errors. We will discuss this further below. Diode pairs D1 & D2 protect the op amp inputs from overvoltage by limiting the voltage swing at the divider output to ±5.6V or thereabouts.

That covers the DC performance of the attenuator, but we want the divider to operate properly up to 25MHz or more. We know that there will inevitably be some capacitance at the output of the divider. The protection diodes, for example, will contribute about 1.5pF each; the op amp input capacitance will be about the same. There will also be 3pF or 4pF of stray capacitance inherent in the layout.

At 25MHz, this ~10pF of total capacitance will have an impedance of around 630Ω, reducing the divider ratio to something in the order of 1/3500. The incidental capacitance is more-or-less unavoidable, so we potentially have a real problem.

The solution is to deliberately add some capacitance across the upper leg of each divider to reduce its impedance by the same ratio and maintain the attenuation. With a 200:1 divider, we would need an upper leg capacitance 199 times lower than the ~10pF in the lower leg. Clearly, this is impractical.

Instead, we put a small known value of capacitance across the upper leg and add more capacitance across the lower leg to compensate for it.

I selected series pairs of 4.7±0.1pF 1kV NP0 capacitors for the upper legs, to match the high-voltage tolerance of the input resistors. Together, they amount to 2.35pF of capacitance in the upper leg of each divider, requiring 467pF of capacitance in the lower leg to compensate.

This latter capacitance is made up of the ~10pF of incidental and stray capacitance we have already mentioned, plus the parallel combination of 390pF and 27pF fixed capacitors, plus VC1 (12-60pF).

This combination gives us a range of capacitance adjustable from nominally 440pF to 490pF. It is useful to have a range to account for capacitor tolerances and other uncertainties. Moreover, the overall bandwidth we can ultimately achieve will be quite sensitive to perfect frequency compensation.

## The buffer stage

The op amp we use for the buffer stage is critical. It must have high input impedances so as not to load the attenuator, and low bias currents since the input impedance is ~10kΩ. Thus, a FET input op amp is required. It must also have a high large-signal bandwidth, and a common-mode input range of ±2V with ±5V supplies.

I chose the ADA4817, which is expensive at around $15 each, but it fits the bill nicely. It has an input impedance of 500GΩ in parallel with 1.3pF and the input bias current is ±20pA. The large signal bandwidth extends to 200MHz, with 0.1dB gain flatness to 60MHz.

The worst-case offset voltage is ±4mV (which is good for a FET input op amp), and the input common mode voltage range is -4.2V to +2.2V with ±5V rails.

If I only required a gain of one for this stage, I could have simply wired IC1 and IC2 as non-inverting buffers. But since we need the option of a gain of 10, I had to close the feedback loop around each op amp with resistors.

It is a good idea to choose a fairly low value for this resistor as it will form an RC low-pass filter with the op-amp's input capacitance, the effect of which will be to increase the gain of the buffer as the frequency rises, causing unwanted 'peaking' in the frequency response.

When the ×10 range is selected via S1, the parallel combination of the 110Ω and 220Ω resistors is switched in between the two buffer amplifiers' inverting inputs. The resistance values were chosen to give this stage a gain of 10 in this configuration. Consistent with the attenuator, I used 0.1% tolerance resistors for gain-setting.

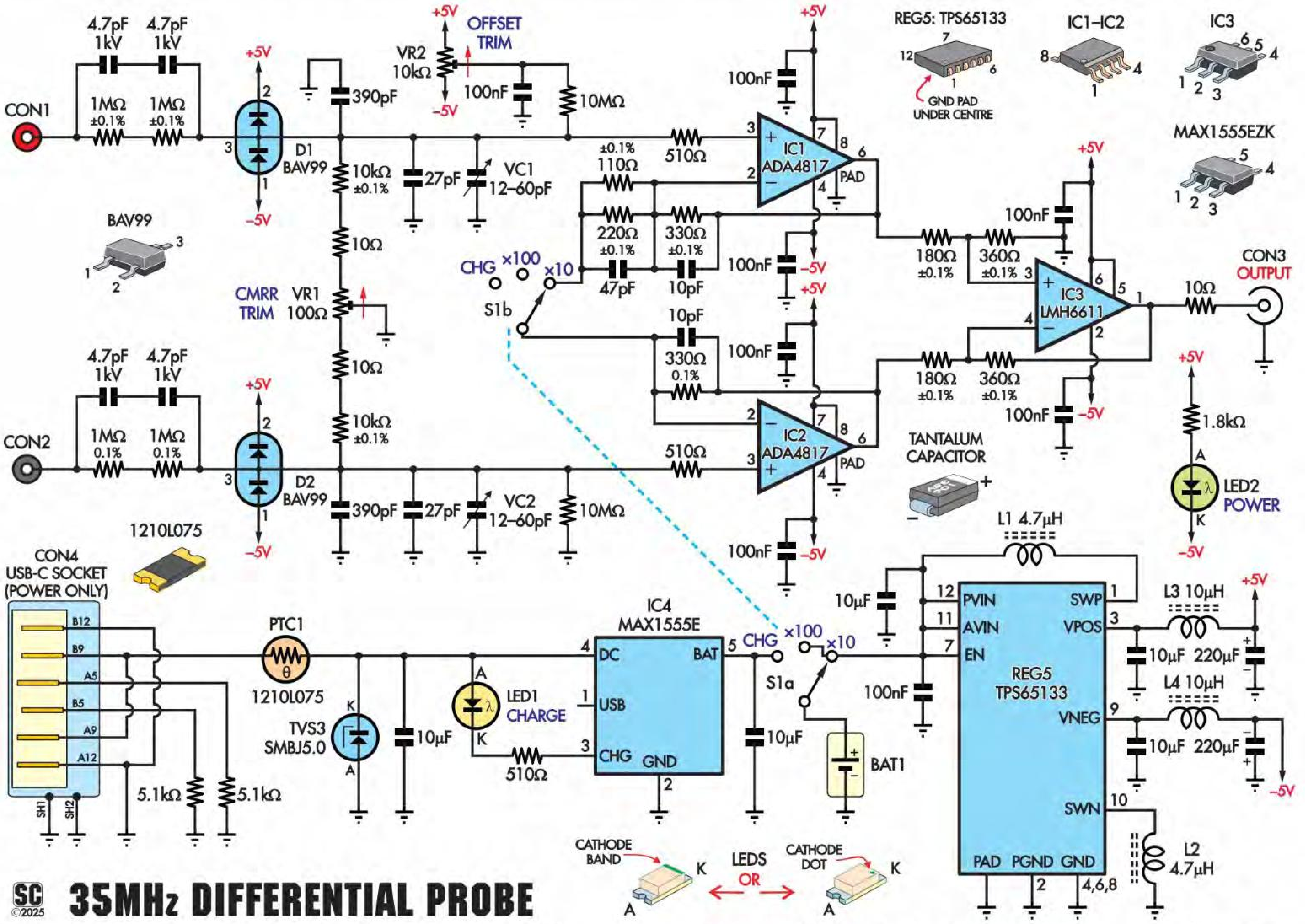
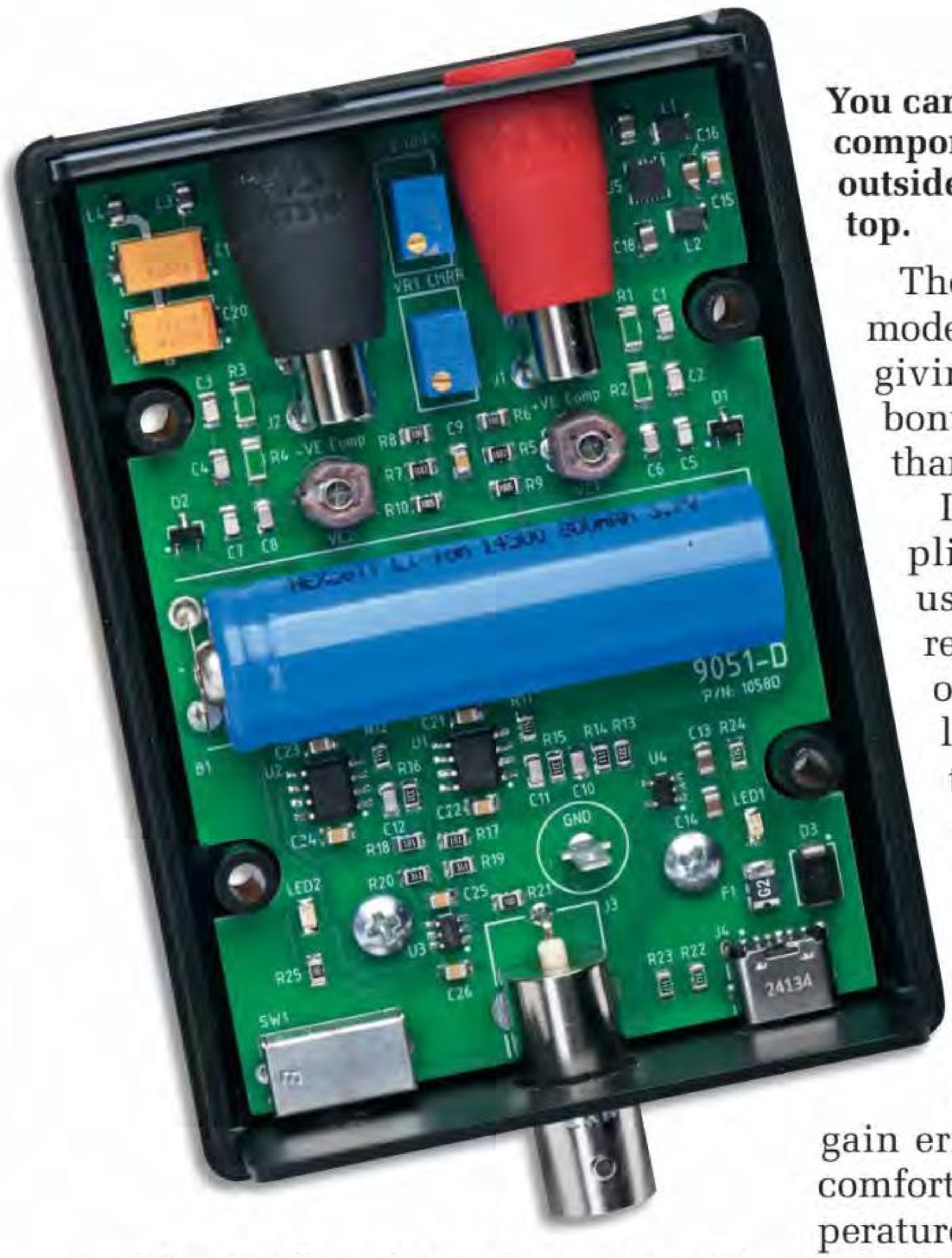

35MHz DIFFERENTIAL PROBE
SC ©2025
CON1
CON2
4.7pF 1kV
1MΩ ±0.1%
D1 BAV99
D2 BAV99
BAV99
+5V
–5V
390pF
10kΩ ±0.1%
27pF
VC1 12–60pF
VC2 12–60pF
10MΩ
10Ω
CMRR TRIM
VR1 100Ω
VR2 10kΩ
OFFSET TRIM
100nF
510Ω
110Ω ±0.1%
220Ω ±0.1%
330Ω ±0.1%
47pF
10pF
CHG ×100 ×10
S1b
IC1 ADA4817
IC2 ADA4817
PAD
180Ω ±0.1%
360Ω ±0.1%
IC3 LMH6611
CON3 OUTPUT
10Ω
1.8kΩ
LED2 POWER
REG5: TPS65133
GND PAD UNDER CENTRE
IC1–IC2
IC3
MAX1555EZK
TANTALUM CAPACITOR
1210L075
CON4 USB-C SOCKET (POWER ONLY)
B12 B9 A5 B5 A9 A12
SH1 SH2
5.1kΩ
PTC1 1210L075
TVS3 SMBJ5.0
10μF
LED1 CHARGE
IC4 MAX1555E
DC USB CHG GND BAT
S1a
BAT1
L1 4.7μH
REG5 TPS65133
PVIN AVIN EN SWP VPOS VNEG SWN PAD PGND GND
L3 10μH
L4 10μH
L2 4.7μH
220μF
CATHODE BAND
CATHODE DOT
LEDS OR

Fig.4: the complete probe circuit. Power is provided by an 800mA Li-ion cell via a dual-rail DC-to-DC converter (REG5). The battery is charged via a USB Type-C connector (CON4) and IC4.



**You can see the input attenuator components arranged vertically outside the banana sockets near the top.**

The 510Ω resistors in series with the non-inverting inputs of IC1 and IC2 are critical to the stability of the circuit. High-speed op amps like the ADA4817 love to oscillate. One of the (many) things that can bring this on is extraneous capacitance on the inputs, and we have plenty given the compensation network we just discussed.

The 510Ω resistors are 'stopper' resistors that isolate the op amp inputs from this capacitance. The 500GΩ input impedance and 1.5pF input capacitance mean that these resistors don't otherwise affect the operation of the probe.

Just as for the input divider, we add 10pF & 47pF frequency compensation capacitors to this gain stage. I did not bother with a variable capacitor here because the low impedance of the surrounding circuit makes it less sensitive to an error of a few picofarads one way or the other.

## Difference amplifier

The requirements for the difference amplifier (IC3) are not quite as stringent as for the buffers, but we do need a high large-signal bandwidth and good output characteristics. The LMH6611 fits the bill. It has a large signal bandwidth of 85MHz and a gain-bandwidth product of 115MHz. The output swing with ±5V rails is ±4.5V into a 150Ω load and the output drive current is ±120mA.

The LMH6611's input common-mode voltage range is -5.2V to +3.8V, giving plenty of headroom. As a bonus, it is considerably cheaper than the ADA4817s.

IC3 is set up as a difference amplifier with a fixed gain of two using low-value 0.1% tolerance resistors. The 10Ω resistor helps overall stability by providing a little bit of isolation between the LMH6611's output and any load capacitance. This stage does not need frequency compensation due to the low gain and low impedances involved.

My design calculations indicate that the end-to-end gain error of this circuit should be comfortably under 1% over the temperature range of 0-40°C, and nearer to half this at 25°C. However, the untrimmed offset error could be in the order of ±5mV on the ×100 range and ±45mV on the ×10 range. The big difference is due to the buffer stage amplifying the ADA4817's offset when on the ×10 range.

This is why it is necessary to add the offset trim. If we added the offset to the difference amplifier (where we would in ideal world), we would need a different offset trimpot for each range and an extra gang on the range switch to select the right one.

The compromise I selected was to add the offset before the gain stage, meaning we can trim out most of IC1's and IC2's offset but may not be able to fully eliminate that from IC3. Since this is ±4mV at the output (0.1% error) in the worst case, I decided I could live with it.

## Power supply

The ±5V power supply is derived from a single Li-ion 14500 (AA-sized) 800mAh cell via a TPS65133 dual-rail switching power supply (REG5). This chip accepts a 2.9-5V input and can source up to 250mA on each rail. It is 92% efficient at 100mA and requires only a couple of inductors and three capacitors to operate.

The chip has an undervoltage lockout to protect the Li-ion cell from over-discharge.

The TPS65133 generates very little noise as far as switching converters go, but to be safe, I added an LC filter (10μH/220μF) between each output of the switcher and the analog circuitry. A green LED (LED2) across the power rails provides user indication that the power is on.

The cell is charged from a USB-C power-only connector via a MAX1555 charger chip (IC4). This charges the cell at around 280mA, which is enough to charge an empty cell in under three hours. A yellow LED (LED1) lights when the MAX1555 is charging the Li-ion cell and extinguishes when it is fully charged.

The charging voltage comes from USB-C connector CON4. Resettable PTC fuse PTC1 and transient voltage suppressor diode TVS3 protect against reverse-polarity circuits and overvoltage conditions. The two 5.1kΩ resistors pull the USB power delivery control channel lines down to passively signal to the source to supply 5V.

The power is switched via a second set of contacts on range switch S1. In the Off/Charge position, the cell is connected to the charger and isolated from the rest of the circuit. In either the 100:1 or 10:1 position, the battery is connected to the switcher and isolated from the charger.

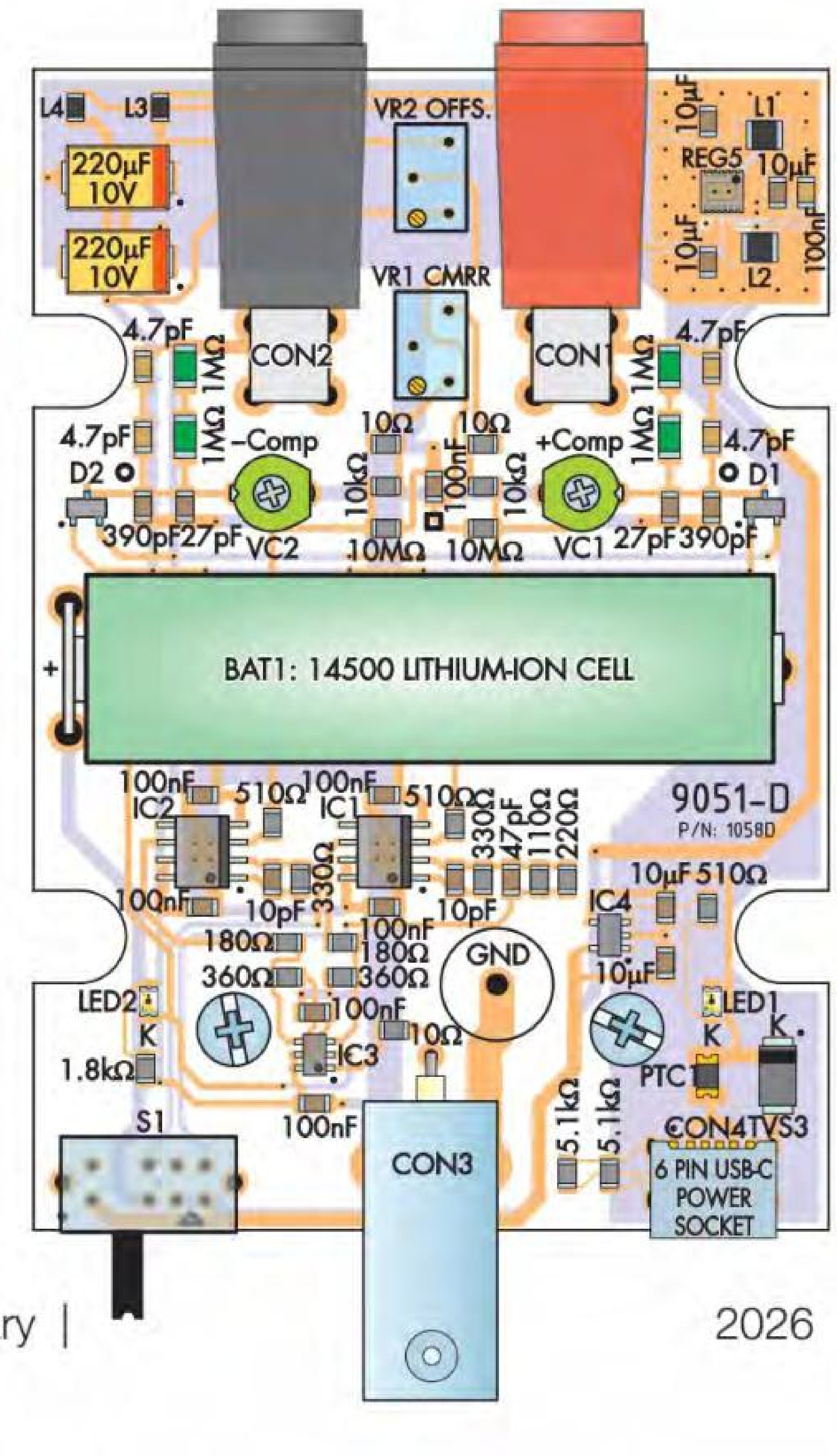


**Fig.5: use this overlay diagram to place the components. We recommend mounting the LCC-packaged DC-DC converter (REG5) and supporting components first. Once you have confirmed they are working, you can move on to the rest of the parts.**

## PCB design

High voltages can be present at the banana sockets' exposed conductors and either end of the first resistor and capacitor of each attenuator. Those components are on the PCB outside the banana sockets.

The track clearances PCB follow IPC2221-B standard B4 for boards with solder masks below 3050m altitude. So you must build this on a commercially-made PCB with a solder mask.

During assembly, you should apply a conformal coating over the top half of the board once all components other than trimpots/trimcaps have been fitted. That will allow it to resist arcing even under extreme conditions (eg, very high humidity).

These coatings are available in spray cans (see the parts list), are easy to apply and can be soldered through, although they should be reapplied later if you do that.

## Construction

All the components mount on a small double-sided PCB coded 9051-D that measures 56.5 × 82.5mm. Most are through-hole or hand-solder-friendly surface-mount types. The only really tricky device is REG5, the TPS65133 switch-mode regulator. Unfortunately, all the useful power chips like it seem to only be available in tiny ‘leadless’ packages.

During construction, refer to the PCB overlay diagram (Fig.5) to see which components mount where and with what orientations.

Because of REG5's package, we recommend assembling and testing the power supply first. REG5 has a thermal pad underneath the chip, so reflow (either hot air or IR) is the only realistic option to mount it. The best way I have found to do this is to use solder paste. Apply a small smear of it to all the pads. Don’t worry if a little gets between pads as it will ball up under surface tension when reflowed.

Place the chip carefully, using the screen-printed lines as a guide. Make sure the orientation is correct. Heat the chip and the surrounding board with hot air until the solder melts, including that on the thermal pad. I use tweezers to hold the chip in place until I feel the surface tension of the solder ‘pull’ it into place.

You can tell the thermal pad solder has melted if the chip re-aligns itself if you nudge it very slightly out of position. Once it cools down, remove any excess solder or obvious shorts with solder wick around the edges (adding a bit of flux paste [not solder paste] makes the wick work better). Then clean up the flux residue with isopropyl alcohol.

Next, fit the four inductors, L1–L4, the four capacitors around REG5 and the two large tantalum capacitors in the upper-left corner of the PCB according to the overlay. You are then ready to test the power supply.

Solder a couple of lengths of fine hookup wire to the board to power it

### Parts List – High-Bandwidth Differential Probe

1 double-sided PCB coded 9051-D, with solder mask, 56.5 × 82.5mm
1 Hammond 1593LBK 92 × 66mm case **[Farnell 4437858]**
1 adhesive panel label, 55 × 80mm
1 14500 (AA-size) Li-ion cell with solder lugs (BAT1) **[eBay 357385477329]**
1 red PCB-mount banana socket (CON1) **[Cal Test CT3151SP-2]**
1 black PCB-mount banana socket (CON2) **[Cal Test CT3151SP-0]**
1 PCB-mount BNC socket (CON3) **[Molex 73100-0105; Farnell 3865358]**
1 USB-C power only socket (CON4) **[Molex 217175-0001; Farnell 3702969]**
2 4.7μH 1.1A M2520/1008 shielded ferrite inductors (L1, L2) **[Würth 74404024047; Farnell 2431471]**
2 10μH 350mA M2012/0805 shielded ferrite inductors (L3, L4) **[TDK MLZ2012M100WT000; Farnell 2215650]**
1 0.75A 24V M3226/1210 PTC polyfuse (PTC1) **[Littelfuse 1210L075/24PR]**
1 right-angle DP3T PCB-mount slide switch (S1) **[E-Switch EG2310*]**
1 top-adjust 100Ω 3296-style multi-turn trimpot (VR1) **[Farnell 9353160]**
1 top-adjust 10kΩ 3296-style multi-turn trimpot (VR2) **[Farnell 9353186]**
2 0.6in (15.24mm) convex light pipes **[Dialight 51513020600F]**
2 No.4 × 6mm self-tapping screws
4 small self-adhesive rubber feet
1 can of low-leakage/high-resistance conformal coating

**Semiconductors**
2 ADA4817-1ARDZ-R7 410MHz precision op amp, SOIC-8-EP (IC1, IC2)
1 LMH6611MK/NOPB 135MHz precision op amp, TSOT-23-6 (IC3)
1 MAX1555EZK-T Li-ion battery charger, TSOT-23-5 (IC4)
1 TPS65133DPDR dual DC-DC converter, WSON-12 (REG5)
1 SMBJ5.0C 5V transient voltage suppressor, DO-214AA (TVS3)
1 SMD M2012/0805 yellow LED (LED1)
1 SMD M2012/0805 green LED (LED2)
2 BAV99 dual series signal diodes, SOT-23 (D1, D2)

**Capacitors** *(all SMD M2012/0805 size 50V NP0/C0G ceramic unless noted)*
2 220μF 10V solid tantalum, SMC case
5 10μF 16V X7R
8 100nF X7R
2 390pF
1 47pF
2 27pF
2 10pF
4 4.7pF 1kV
2 6mm diameter 12-60pF variable capacitors (VC1, VC2) **[EW GKG60015]**

**Resistors** *(all SMD M2012/0805 size ±1% ⅛W unless noted)*
2 10MΩ
4 1MΩ ±0.1% M3216/1206 size ¼W 700V **[Vishay TNPV12061M00BEEN]**
2 10kΩ ±0.1% 10ppm **[Farnell 1140912]**
2 5.1kΩ
1 1.8kΩ
3 510Ω
2 360Ω ±0.1% 25ppm **[Panasonic ERA-6AEB361V; Farnell 1670222]**
2 330Ω ±0.1% 25ppm **[Panasonic ERA-6AEB331V; Farnell 4137016]**
1 220Ω ±0.1% 25ppm **[Panasonic ERA-6AEB221V; Farnell 1577649]**
2 180Ω ±0.1% 25ppm **[Panasonic ERA-6AEB181V; Farnell 4756914]**
1 110Ω ±0.1% 25ppm **[Panasonic ERA-6AEB111V; Farnell 1670214]**
3 10Ω

*** Farnell 2435089 is pin-compatible but has a 2mm shorter actuator, so it might not project fully through the end panel.**

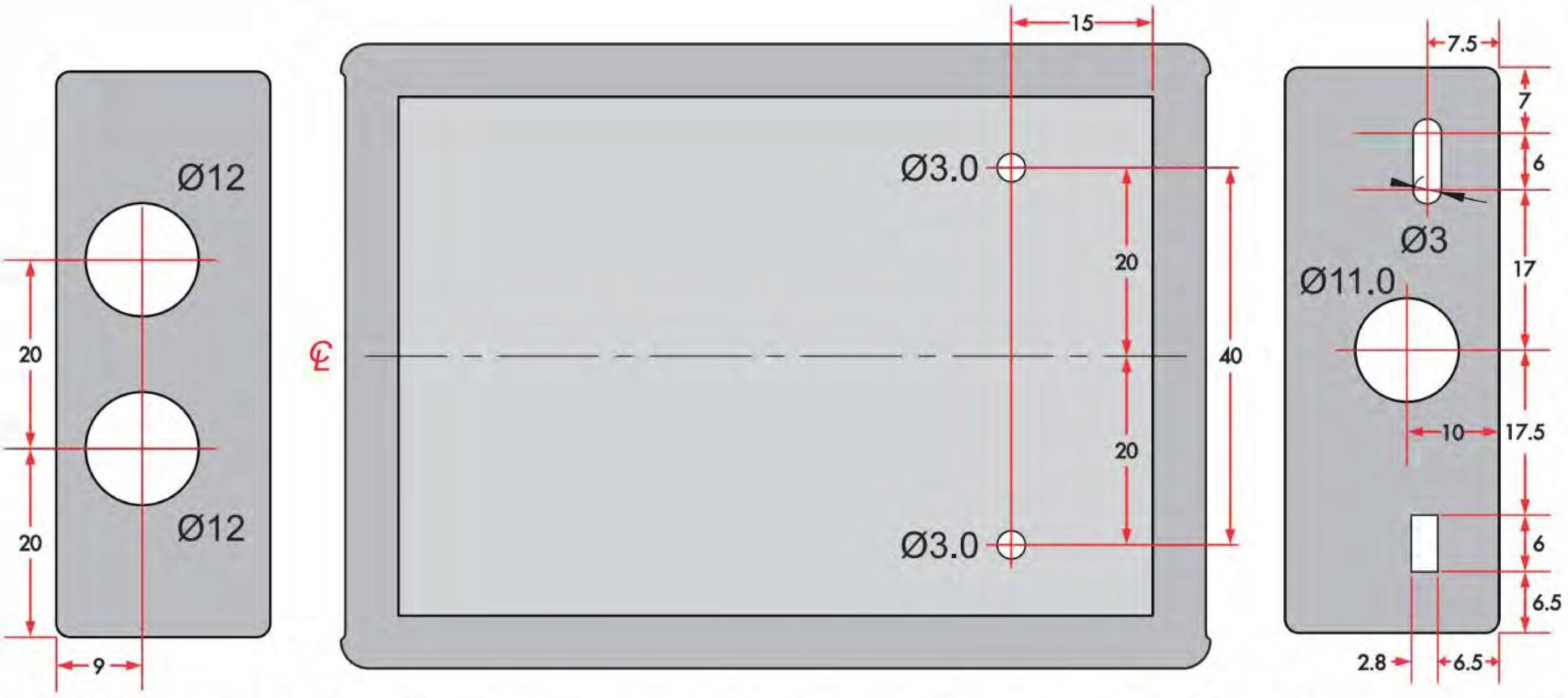


**Fig.6: drill the enclosure end panels and top according to this diagram. The slots can be formed by drilling a pair of holes inside the perimeter and using a craft knife and files to open them up to the required dimensions.**

externally. The easiest place to connect the negative supply is the through-hole for the battery negative terminal (the single hole on the right-hand side of the board). The best place to connect the positive supply is the bottom right-hand through-hole in the group of six where the switch will later be mounted.

These locations are marked by small triangles on the PCB silk screen overlay.

Connect an external power supply set to deliver 4V with a current limit of 100mA and switch it on. The current draw should be negligible, and you should be able to measure 5V across both of the large tantalum capacitors. If there is a problem, switch off, check your work and, if necessary, reflow REG5 again.

If all is well, you can remove the wires and proceed with mounting all the other parts, leaving the battery till the very last. IC1 and IC2 also have thermal pads on the bottom, so these will have to be reflowed too. However, they are SOIC-8 packages so are much easier to solder than REG5.

Remember to apply the conformal coating we mentioned earlier on both sides of the board above the cell location before soldering the trimpots and trimcaps. Reapply it on the underside after soldering those components so their joints are covered.

The cell used on the prototype has PCB pins, but that configuration is hard to find, so we have specified a type with small solder tabs that can be attached using short lengths of stiff wire. Ensure it is orientated correctly before soldering the wires!

Immediately after you mount the cell, screw the board into the case bottom. This will help prevent accidental shorts under the board. The energy density of the Li-ion cell is such that accidental shorts can easily burn out tracks or cause other damage.

## Case preparation

Drill the enclosure end plates and top case according to Fig.6. The slots can be most easily made by drilling a couple of holes inside the outline and finishing with a craft knife and small files. You will need to remove the two plastic bosses on the inside of the top case where the banana jacks are located – you can just snip them out with a pair of side cutters.

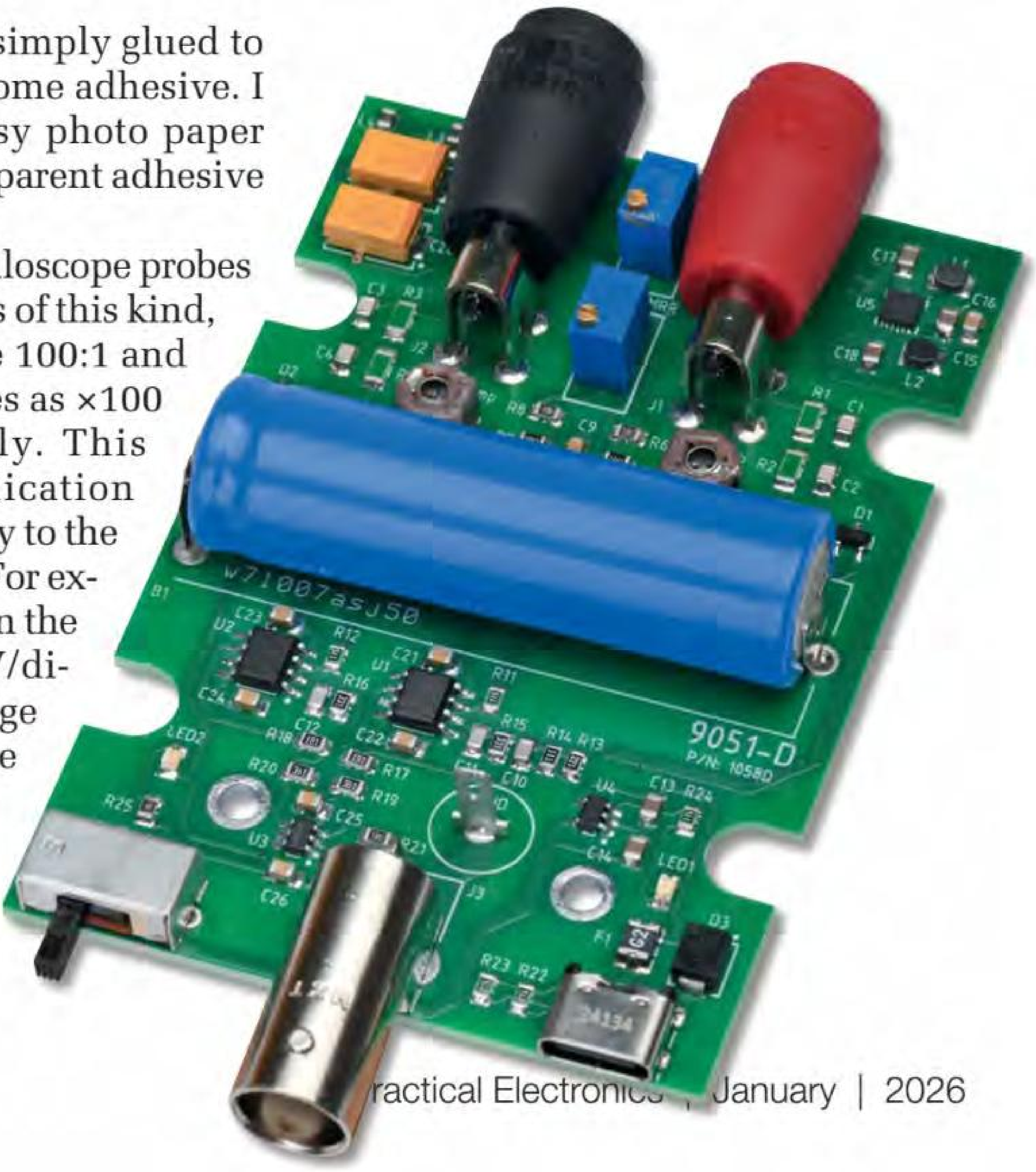
The label (Fig.7) is simply glued to the front panel with some adhesive. I printed mine on glossy photo paper and covered it in transparent adhesive film for protection.

Consistent with oscilloscope probes and commercial probes of this kind, the label describes the 100:1 and 10:1 attenuation ranges as ×100 and ×10, respectively. This refers to the multiplication factor you need to apply to the 'scope's vertical scale. For example, a 1V/division on the scope represents 100V/division on the ×100 range and 10V/division on the ×10 range.

Once it has been applied, punch out the holes for the light pipes and push them in from the front. They can be secured with a drop or two of cyanoacrylate adhesive (superglue) on the back side. Assemble everything except the top case and you are ready for calibration.

## Testing and calibration

Start the calibration process by fully charging the battery. Connect a USB-C power supply to the probe and make sure the switch is in the off/charge position. The yellow LED should light, indicating the battery is charging. When full charge is reached, the LED will go out. This may take two or three hours if the battery is nearly discharged.



**The assembled PCB before it was installed in the case.**

Once charged, remove the USB cable and power the unit on by selecting either the ×100 or ×10 range and recheck that the power supplies are at ±5V as before. The green LED should be lit.

The first step in calibration is to zero out the offset correction. We need to do this to make sure it does not impact the setting of the CMRR trim in the next step. Switch the probe to the 100:1 range and adjust VR2 until the voltage at its wiper is as close to zero as you can get it.

You can clip your voltmeter's negative lead to the GND test point and read the wiper voltage on the bottom end of the vertically orientated capacitor immediately below VR1, marked by a small square on the PCB overlay. You should be able to adjust the voltage to within a few millivolts either side of zero. Anything under ±10mV is fine.

Now we need to adjust the CMRR trim. Set your bench power supply to the highest (safe) voltage you can get. For example, connect two channels of a dual 30V supply in series for 60V. Connect the positive lead of the power supply to both of the probe inputs (shorted together) and the negative lead to the GND test point. Switch the probe to the 100:1 range.

Use your meter to measure the voltage between the mid-points of the voltage dividers while you adjust VR1. The suggested probe points are marked by small circles on the PCB overlay, immediately to the left of D1 and the right of D2. Adjust VR1 for a reading as close to zero as you can get at these points. You should be able to get a reading below ±20µV.

With a 60V input, a reading below ±20µV implies a CMRR of 130dB. But you can probably do better than that with a good meter and some patience.

Now you can set the offset voltage trim. Remove the power supply but keep the two inputs shorted. Measure the output voltage at the BNC connector with respect to the ground test point. Trim VR2 to get the output close to zero on both ranges. This may require a little backwards and forwards between ranges and the acceptance of some compromise (for reasons described above).

For example, the best I could do was -1.1mV on the 100:1 range and +1.5mV on the 10:1 range. You should be able to get to within ±10mV of zero on both ranges simultaneously.

The final step is to trim the frequency compensation. You will need a function generator and an oscilloscope. The function generator should be set to deliver a 1kHz square wave at the highest amplitude you can manage. Connect the differential probe to the scope using a BNC-to-BNC cable and make sure the scope's bandwidth limit is disabled.

To set up the positive divider, connect the function generator's output to the positive input of the probe and its common to the ground test point. Also connect the probe's negative terminal to the ground test point. Switch the probe to the 100:1 range.

Set up your scope to get a stable display of the square wave output of the differential probe and adjust compensation trimmer VC1 for optimum compensation, just as you would for an oscilloscope probe. The correct compensation is achieved when the rising edge of the square wave shows no overshoot or undershoot, as shown in Fig.8.

Use a non-metallic tool to make this adjustment. It's better to err very slightly on the side of over-compensation (a small amount of overshoot) if you are unsure, as this will maximise the probe's bandwidth.

Repeat the whole process for the negative divider, connecting the function generator output to the negative input of the differential probe and the probe's positive terminal to the ground test point. This time, tweak VC2 for optimum compensation.

## Using it

Screw the lid on and your probe is ready to use. I added four small self-adhesive rubber feet to the bottom of the case to prevent it from sliding around too much on the bench.

Always take special care when you are using the probe with high voltage circuits. Make all connections – including that from the probe to the scope – before powering up any circuit under test. Never disconnect any high-voltage differential probe from the scope while the test circuit is powered on. If you do, the BNC connector on the probe can float to high voltages.

There is no isolation barrier in these devices. Not much current can flow due to the high impedance of the probe, but you can still get a shock. Always use quality test leads with shrouded banana plugs for high-voltage connections, and check everything twice before powering it up. PE

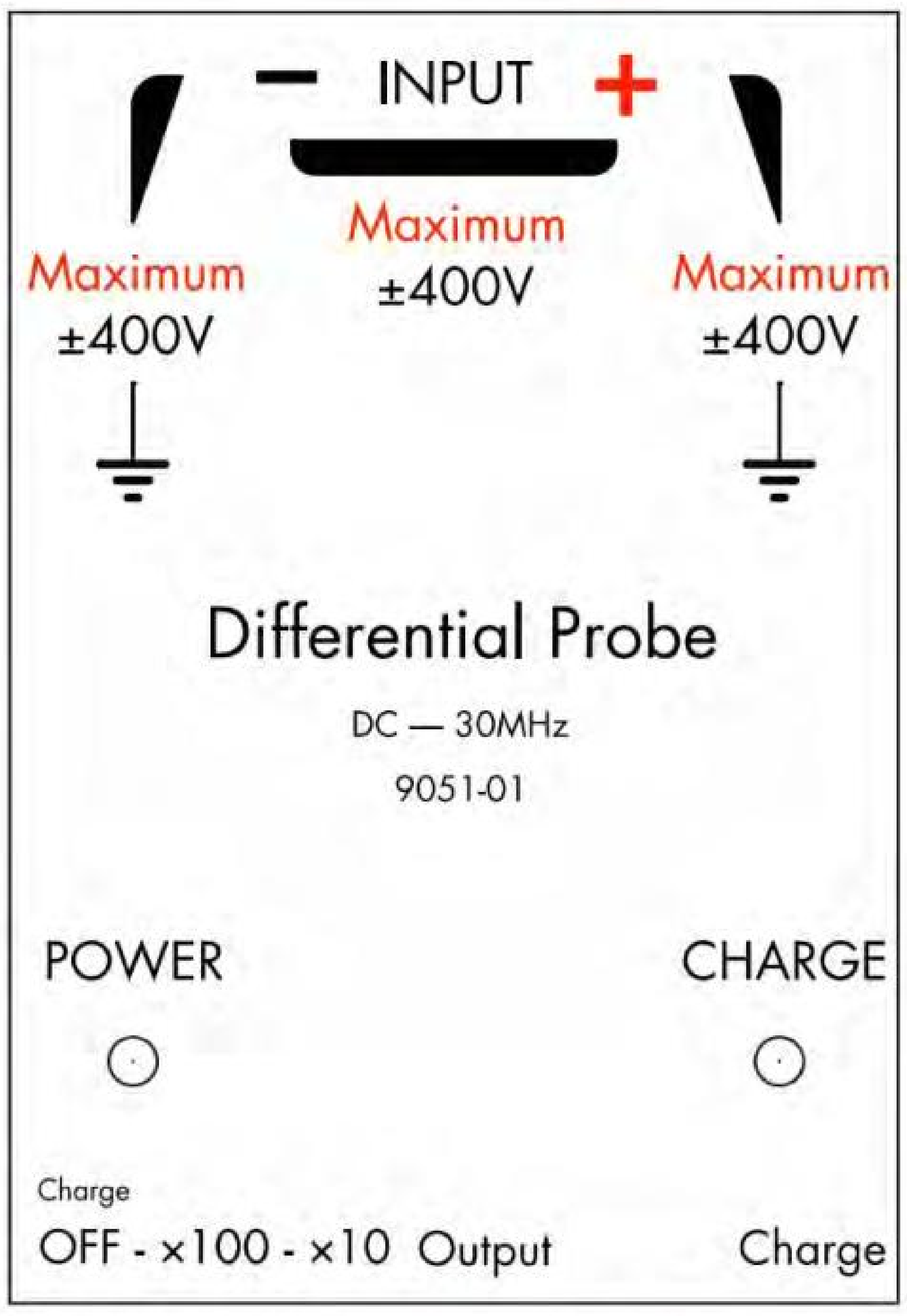
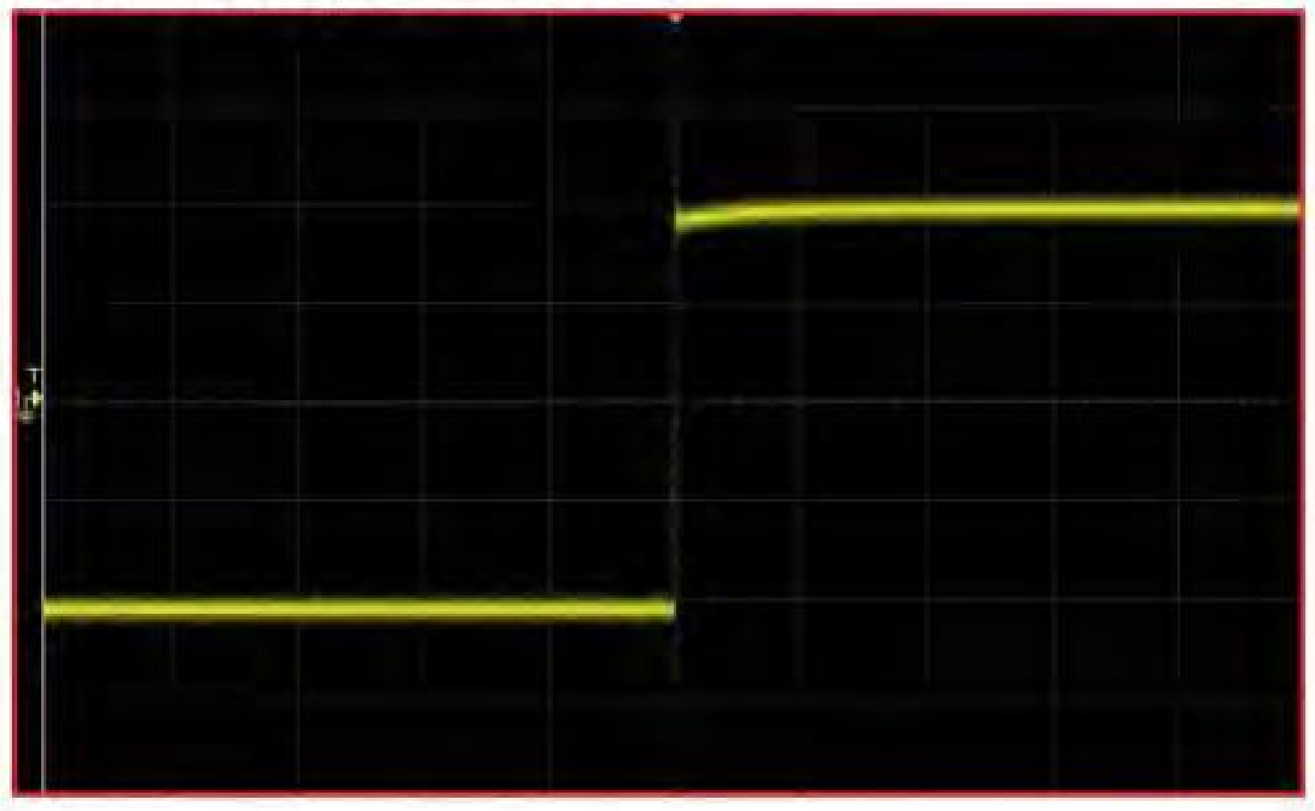
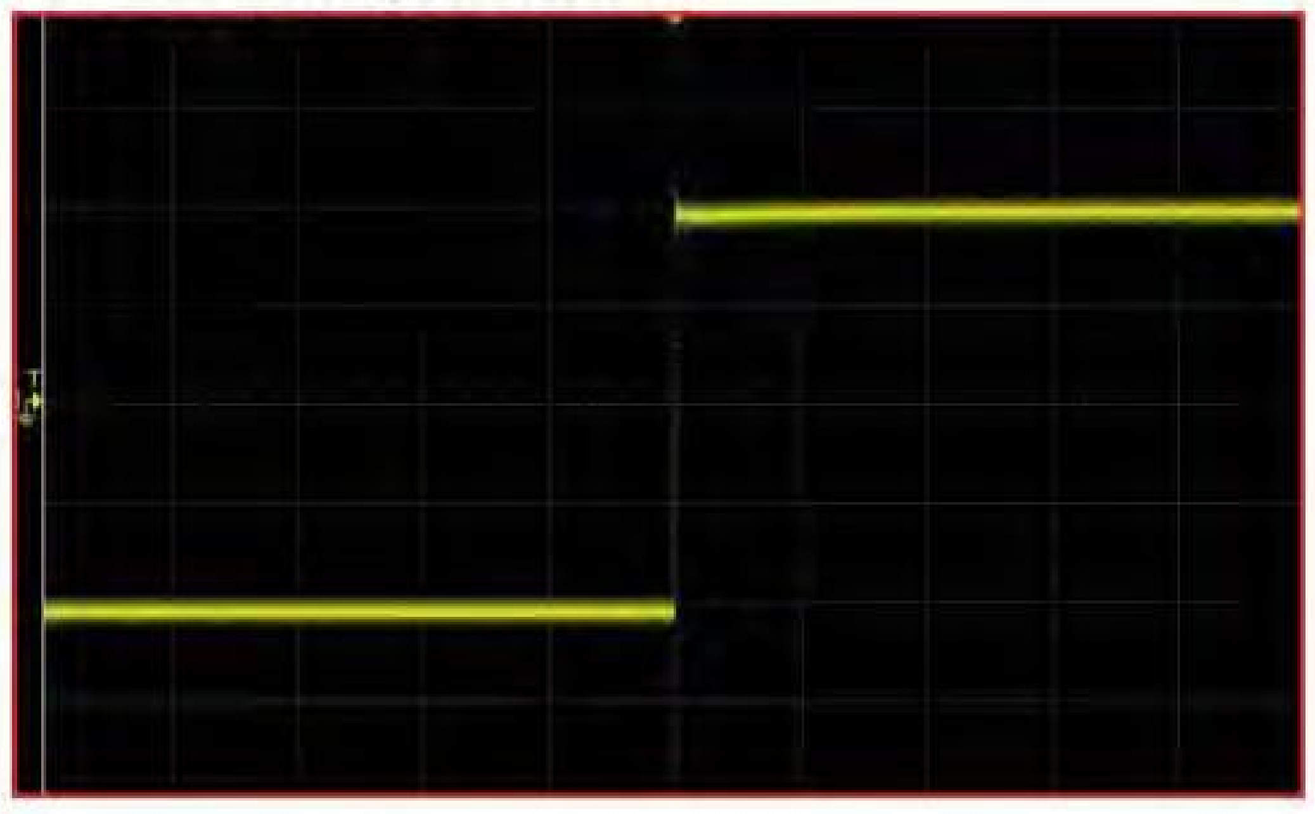


Fig.7: this label artwork can be downloaded from https://siliconchip.au/Shop/11/607 as a PDF. For details on how we make front panels see siliconchip.com.au/Help/FrontPanels

**Undercompensated**



**Correct Compensation**



**Overcompensated**

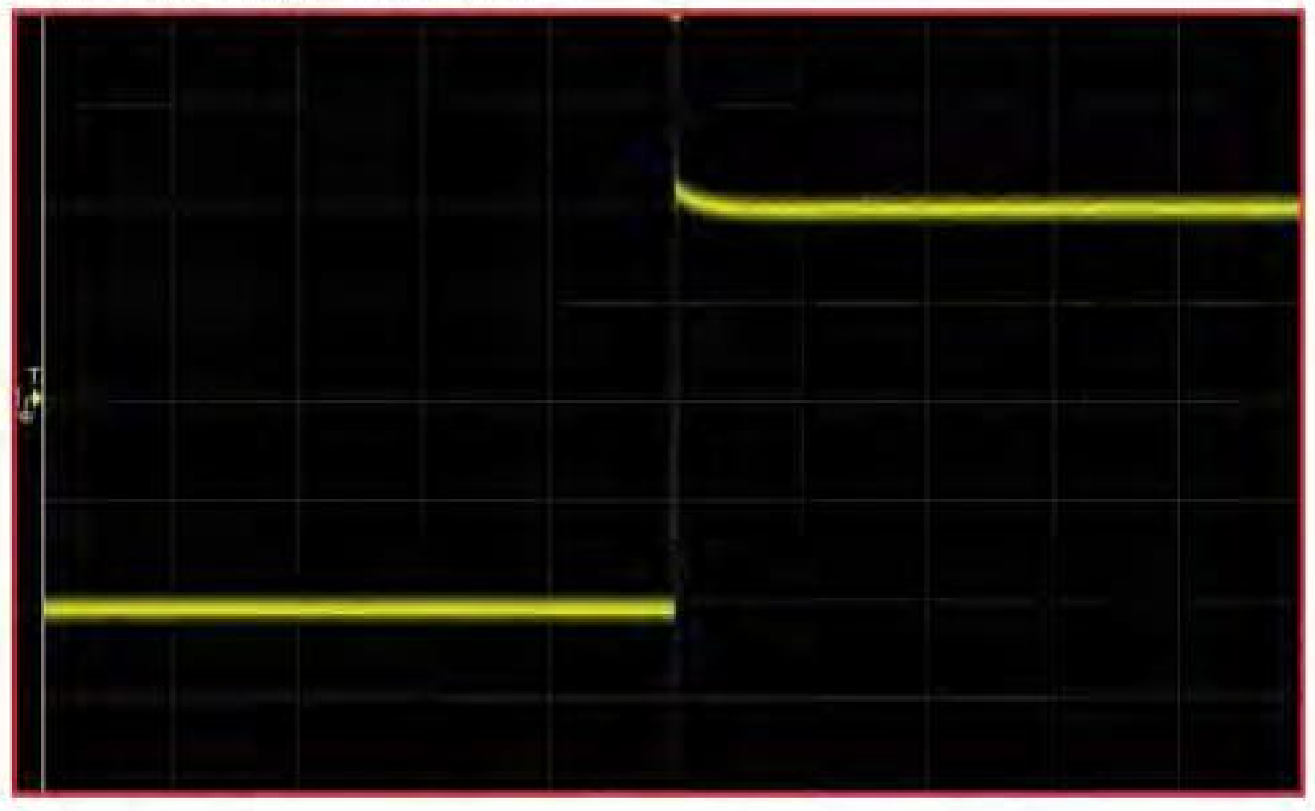


Fig.8: correct compensation is achieved when the square wave's leading edge shows no undercompensation droop or overcompensation overshoot.

# Thinking without power

Techno Talk

Max the Magnificent

From crawl-space conundrums to battery-free Bluetooth, join me as I explore how tomorrow's AI-powered sensors might think, sense and communicate while barely sipping power, or even skipping it altogether.

**I feel that we're poised at the** edge, just around the corner from an inflection point. In the not-so-distant future, we'll be submerged in a metaphorical sea of ultra-low-power connected, AI-enabled sensors.

In addition to granting us access to vast amounts of information about the world around us, these systems will be able to tell us not only when something has failed but also when it's about to fail. Unfortunately, we aren't there yet (said Max, crankily).

### Smart homes? I should cocoa!

The catalyst for my current crankiness is that we—my wife, Gigi the Gorgeous, and your humble narrator—recently discovered a moisture problem in the crawl space beneath our house. I currently hang my hat in Huntsville, Alabama, USA, which "enjoys" a hot, damp and humid climate.

It turns out that the humidity in our crawl space is off the scale. If left unchecked, it would eventually lead to mould, which could compromise the structural integrity of the wooden joists and floors above.

As I pen this column, I can hear a bunch of guys working under the house. They are in the process of encapsulating the crawl space. This involves laying a thick, puncture-resistant vapour barrier across the entire crawlspace floor and up the sides of the walls (any seams are taped or heat-sealed).

Next, a layer of insulation is added. The insulation we're using is a gleaming, silver-foil-faced reinforced bubble sheeting that will make our crawl space look like the inside of a spaceship. Last but not least, they will add an industrial-strength dehumidifier that will run continuously to suck moisture out of the air and expel it into the outside world.

Nothing runs forever. How will I know if my new (very expensive) dehumidifier starts to fail? Is it perhaps equipped with a battery of state-of-the-art sensors (eg, temperature, pressure, current, voltage, vibration), coupled with AI and sensor fusion, that can detect the onset of problems and alert me via a Wi-Fi link?

### I should cocoa!

Apparently, I need to periodically open the entrance hatch and crawl across the aptly named crawl space to physically inspect the unit, which will be positioned under the centre of the house.

Suffice it to say that I'm wearing my frowny face. I'm also wearing my "I think I may have an idea for a new project" face.

### Miniature meets mighty

As I've mentioned in previous columns, one of the great aspects of my role as a freelance technology consultant is that I get to chat with a diverse range of people about a wide range of things.

For example, I recently spoke with Josh Elijah. Josh loves miniaturising things, and he's obsessed with creating tiny electronics products. A couple of years ago, he turned his attention to designing miniature Ethernet routers. People loved these products so much that Josh co-founded BotBlox (**botblox.io**) with his brother Aaron, and the company has since grown into a multi-million-dollar business.

Josh's "latest and greatest" creation is an 8-port, 10/100Mbps Ethernet switch and router called SwitchCore. At just 1" × 1" × 0.5" (~25 × 25 × 12mm), this little scamp is ideal for many weight-, power- and volume-constrained applications ranging from drones to space probes.

### From prompt to prototype

Do you design your own electronic projects? If so, do you also design your own printed circuit boards (PCBs)? If you enjoy doing this, more power to you! If not, wouldn't it be great if you could simply state your requirements and have the project designed and built for you?

To pull a random example out of thin air, I might like to say something like: "I have a dehumidifier under my house. I want to know if it stops working without having to crawl under there. Can you design a little gadget that plugs into the wall, senses vibration from the dehumidifier, and sends me a Wi-Fi alert if it hasn't been running for a while? It needs to survive a humid crawlspace and be safe on household mains power."

Luckily, in my case, I already have this covered. My chum Joe Farr loves challenges like this. If I call him up on a Friday to pose a problem like this, he will have a board designed and on its way to being fabricated before the weekend is out.

But what about people who can't do this themselves and don't have their own Joe on speed dial? Well, I recently spoke with Matthias Wagner, founder and CEO of Flux (**flux.ai**). Matthias tells me that the latest version of Flux can accept a prompt like my "I have a dehumidifier under my house…" Flux can ask supplementary questions if necessary.

Once it has wrapped its AI brain around your requirements, Flux creates and deploys AI agents that perform its bidding, which includes designing the circuit, generating the schematic, selecting components and laying out the PCB. The only thing Flux doesn't currently do is order the boards for you, but Matthias says the team is working on that.

### Battery-powered AI at the edge

In the case of my hypothetical dehumidifier monitor, power isn't much of a consideration because I can ride on the back of the mains power that's already supplying the machine. But what about battery-powered AI applications? I'm glad you asked.

EMASS-I2R
MOSCHIP
ECS-DOT
2231
Singapore

The ECS-DoT chip is shown here six times wider than in reality. It's actually only 10 × 10mm (roughly 3/8" × 3/8").

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A few weeks ago, I had the opportunity to chat with Mark Goranson, the CEO of a fabless (and fabulous!) semiconductor company called EMASS (**https://nanoveu.com/emass/**), based in Singapore.

The folks at EMASS have developed a device called ECS-DoT, which stands for "Edge Compute Subsystem—Deep Optimized Tensor". The evaluation incarnation is presented in a 10 × 10mm package; the production version will be available in a 5 × 5mm package, and a 2 × 3mm chip-scale package (CSP) version will be made available in the future.

Mark says that, compared with its leading competitors, the ECS-DoT operates up to 93% faster while consuming 90% less energy, all while supporting true multi-modal sensor fusion on-device. In real terms, the ECS-DoT can perform 30 giga operations per second (GOPS) while consuming a paltry two milliwatts.

### But what about the radio?

Quite a few companies are boasting about their low-power Edge AI processors these days, particularly in the context of small, portable, or remote devices that operate on limited power sources such as batteries or using energy harvesting.

Unless those are a giant's fingers, this Ethernet switch/router is tiny!

What's often "brushed under the carpet", as it were, is the matter of connectivity—especially wireless links like Bluetooth or Wi-Fi. In many IoT applications, the wireless radio link can dominate the energy budget. The dirty little secret is that communicating data can burn far more energy than thinking about it.

For example, the processor might sip a few tens of microwatts while running an inference, but the radio can gulp tens or even hundreds of milliwatts just to say, "Hello, world!".

Recently, I heard about an interesting company called Haila (**haila.io**). Yes, of course, I ended up chatting with them.

Their 'claim to fame' is extremely low-power Bluetooth and Wi-Fi for the Ambient IoT, which refers to a new class of ultra-low-power or battery-free connected devices that can sense, compute and communicate by harvesting energy from their environment.

The explanation they provided to me is as follows.

*If you use a mirror to reflect the light from a laser, this doesn't require you to consume power (other than a tiny bit to move the mirror). Similarly, instead of actively transmitting a high-power RF signal, a Haila device can reflect or modulate an existing RF signal (or harvest ambient RF) to communicate.* As a result, they boast 1000 times lower Wi-Fi and Bluetooth power consumption compared to conventional solutions.

I commenced this column talking about "a metaphorical sea of ultra-low-power connected, AI-enabled sensors". Technologies like those from EMASS and Halia may make these devices real before we know it! **PE**

## MAX'S COOL BEANS

**By Max the Magnificent**

# WEIRD & WONDERFUL ARDUINO PROJECTS

## Part 13: numerical superiority & defying logic

**Goodness gracious me; I simply** cannot believe it's January 2026 already. I haven't prepared a speech, and I've nothing appropriate to wear!

For those who are joining us for the first time, we are slowly but surely creating a retro games console. My chum, Joe Farr, already has a prototype up and running, as shown in Photo 1.

A bird's-eye, not-to-scale view of the console's front panel is shown in Fig.1. As we previously discussed, you can fabricate your console's front panel using a laser cutter. Also, you can use a 3D printer to construct the case.

If neither you nor your friends has a laser cutter or a 3D printer, you can always join a local makerspace (also known as a hack space, hacker space, fab lab or open workshop). Alternatively, you can start off by simply attaching everything to a piece of pressed board or ¼-inch plywood (remember that hot glue is your friend, unless it gets on your skin).

We commenced this epic extravaganza by implementing the 14 × 10 array of tricolour LEDs that will occupy—nay, dominate—the central area of the front panel. This will form our main play area.

Next, we introduced the concept of switch bounce and evolved a cunning scheme to debounce switches in software. In the fullness of time, we will utilise a modified version of this software to debounce and access the state of the eight switches at the bottom of the fascia panel (shown as "Direction Controls" and "Action Controls" in Fig.1).

More recently, we moved to consider the 7-segment displays at the top of the front panels. We will be employing these little scamps to display information such as the current player (in multi-player games), the current level (in multi-level games) and the current score.

We're utilising prebuilt modules that include the displays, current-limiting resistors and 74HC595 8-bit serial-input, parallel-output (SIPO) registers in one handy-dandy package. These modules come with two, three or four digits. Like Joe, I'm using three two-digit modules, but you can use as many (or as few) as you wish.

Now let's plunge headfirst into the fray with gusto and abandon.

### Installing the display modules

In our November 2025 column, we performed some experiments to determine which segments of the 7-segment displays on our modules are connected to which bits (outputs) from their 8-bit shift registers.

As part of this, much to our (well, my) surprise, we discovered that my modules feature common anode (CA) displays, as opposed to the common cathode (CC) components we'd been expecting. We quickly developed a solution to address this issue in our software.

In last month's column, we decided on a preliminary set of characters we wish to display: the digits 0 through 9, space (blank), hyphen (minus), "P" for "Player" (in multi-player games), "L" for "Level" (in multi-level games), along with "E" and "F" for "Error" and "Fail", respectively.

We also created a few useful functions: one to modify a score with a positive or negative delta (change) value while staying within minimum and maximum bounds; one to load a numerical value into our display buffer using a specified number of digits with a defined offset (from the least significant, rightmost digit); and one to copy the contents of the display buffer to the displays themselves.

In all these early experiments, we were working with only a single two-digit display module. Now it's time to connect multiple modules together, as illustrated in Fig.2.

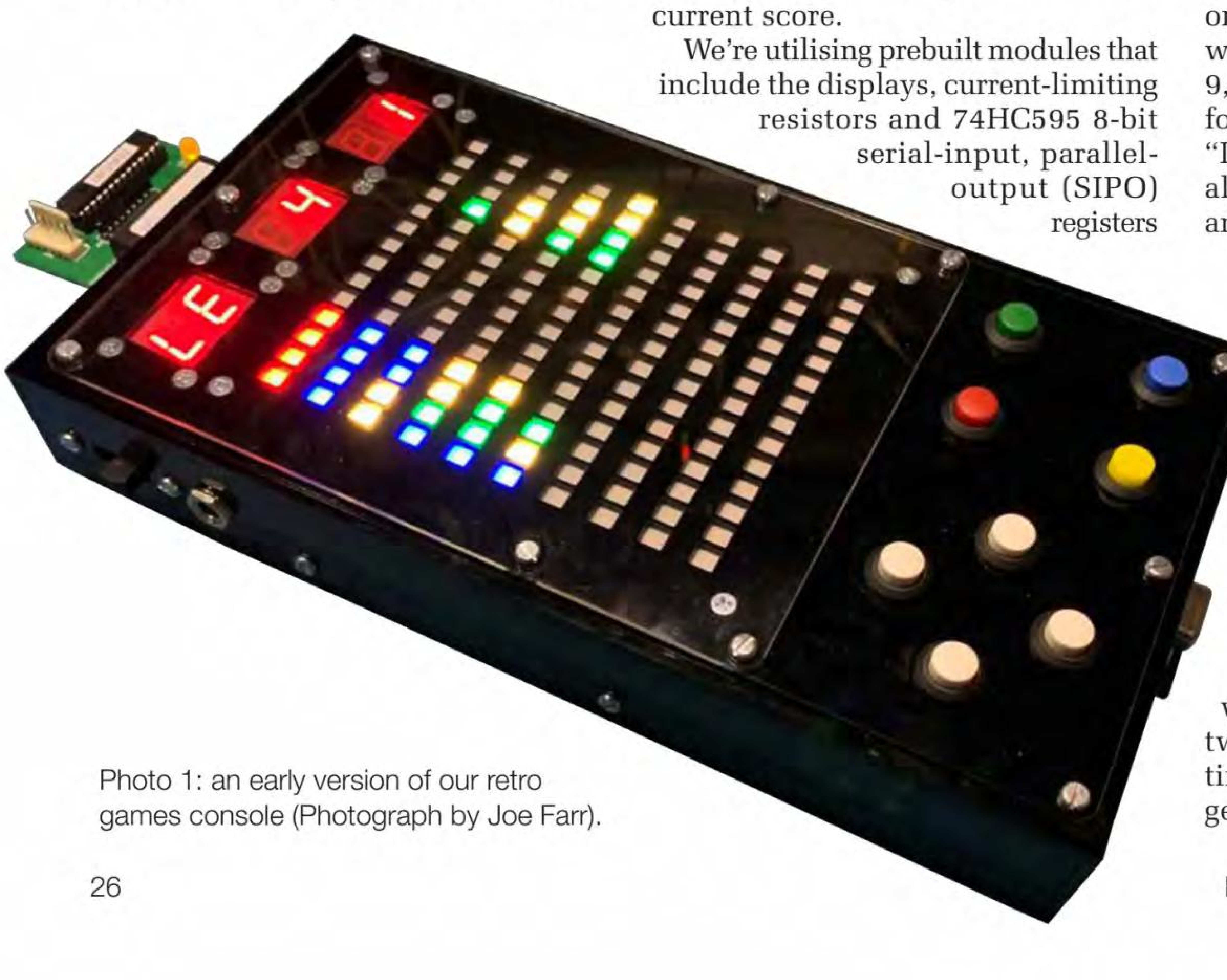


Photo 1: an early version of our retro games console (Photograph by Joe Farr).

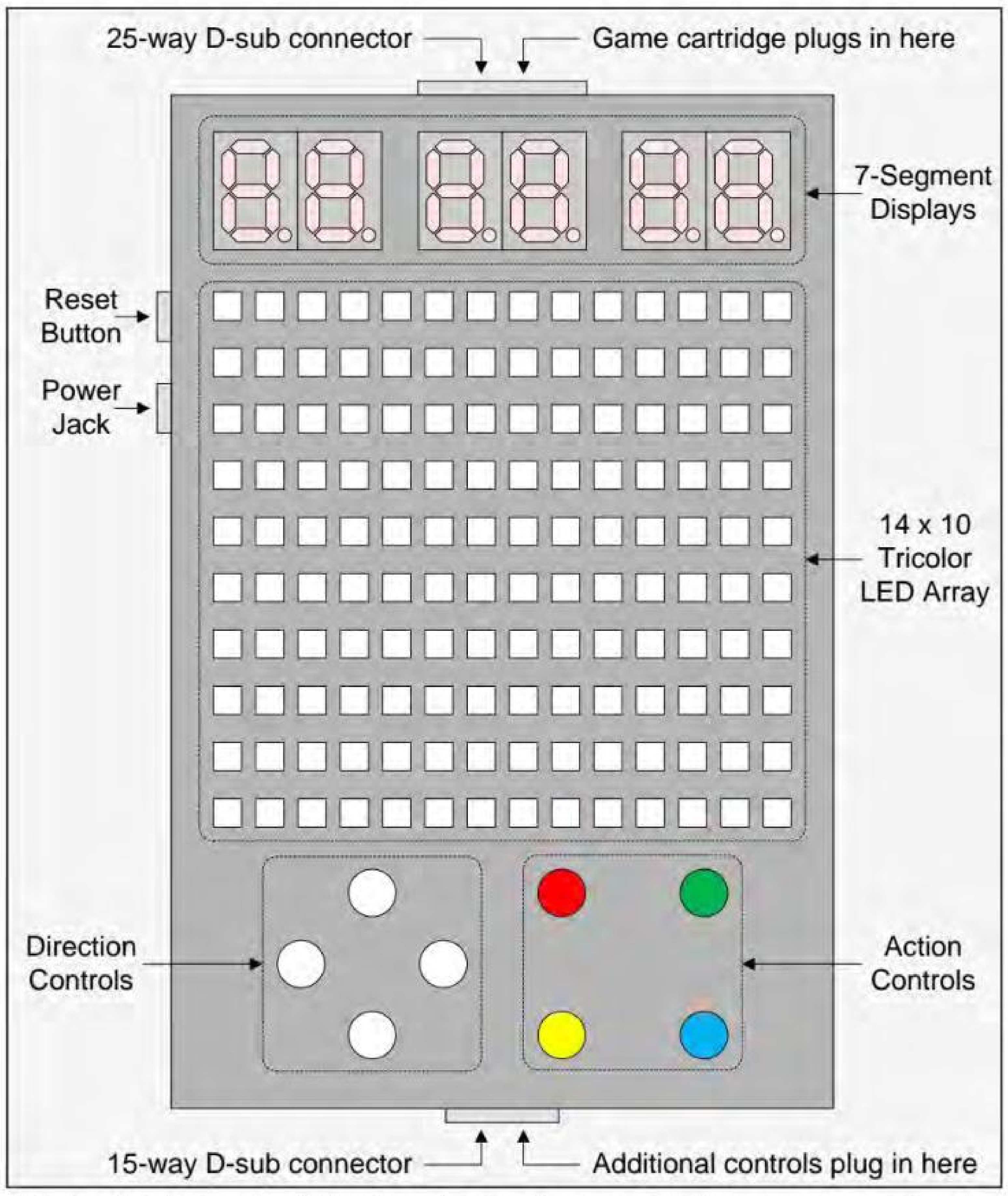

25-way D-sub connector
Game cartridge plugs in here
7-Segment Displays
Reset Button
Power Jack
14 x 10 Tricolor LED Array
Direction Controls
Action Controls
15-way D-sub connector
Additional controls plug in here

Fig.1: the layout of the front panel on our retro games console.

There are multiple ways to achieve this. For example, if I were planning to mass-manufacture these consoles, I'd start by making a jig that would allow me to insert a display module, apply power and run some tests using spring-loaded pogo-pin contacts instead of wired connections. This would allow me to quickly test each module individually before proceeding. However, since I have only three modules to worry about, creating such a jig seems like a lot of effort for a small reward.

From our earlier breadboard-based experiments, I have already soldered a 5-pin header onto the left-hand set of pins on what will henceforth be known as Module 2. However, I've omitted this from Fig.2 to avoid confusion.

We could solder the wires connecting the modules from the front before attaching them to the front panel. In this case, things would appear as shown in Fig.2. For myself, I decided to first attach the modules to the front panel, as illustrated in Fig.3(b) overleaf, then solder the wires from the back, as in Fig.3(d).

**Note** that the "Cutting Plane Line" shown in Fig.3(a) does **not** mean you cut the panel. In technical drawings, a cutting plane line (or section line) indicates the location where an imaginary cut has been made through an object to reveal internal features. This allows us to see how the 7-segment displays are fitted within the openings in the front panel in Fig.3(b).

Regarding the attachment of the modules, I utilised parts from a black M3 PCB spacer kit (pemag.au/link/ac4u). As shown in Figs.3(b) & (c), the 6mm (~¼in) spacers, in conjunction with 12mm (~½in) machine screws and nuts, result in the fronts of the 7-segment displays being flush with the face of the panel.

### Testing the multi-module display

In the spirit of adventure, the first test I performed was to take our latest and greatest program from last month and run it on my new 6-digit display.

For your edification, I've made a copy of this code available in the file named **CB-jan26-code-01.txt** (as usual, all files mentioned in this column are available as part of the January 2026 download package from the PE website at https://pemag.au/link/ac9b).

This program includes a function called **NumberToBuffer()** that accepts three parameters: the number we want to load into our display buffer, the number of digits we wish to display, and the number of digits by which we wish to offset the value we display from the least significant digit (LSD) on the right of the display.

If offset is 0, then the rightmost digit of the value to be displayed will appear on digit 0 of the display; if offset is 1, then the rightmost digit of the value to be displayed will be loaded into digit 1 in our buffer and appear on digit 1 of the display; and so forth.

As you may recall, the current version of this program is based on a 2-digit display, so its **NUM_DIGS** ("number of digits") preprocessor macro constant is set to 2.

The program repeatedly modifies the value of **P1Score** (Player One's score), counting up from **MIN_SCORE** to **MAX_SCORE** and then back down to **MIN_SCORE**, where **MIN_SCORE** and **MAX_SCORE** are 0 and 99, respectively.

The key part of this program (insofar as our discussions here) is where we call the **NumberToBuffer()** function to display the current value of **P1Score** using the following statement, where 2 is the number of digits to be displayed and 0 is the offset.

```
NumberToBuffer(P1Score, 2, 0);
```

Without running the program, what do you think will happen?

Since the program thinks it's writing to a 2-digit display, we end up with the following sequence:

```
ΔΔ ΔΔ ΔΔ
00 ΔΔ ΔΔ
01 00 ΔΔ
02 01 00
03 02 01
04 03 02
 :  :  :
99 98 97
98 99 98
97 98 99
96 97 98
 :  :  :
```

I'm using triangle (**Δ**) characters to indicate blanks (all segments off) on the displays. If things are happening

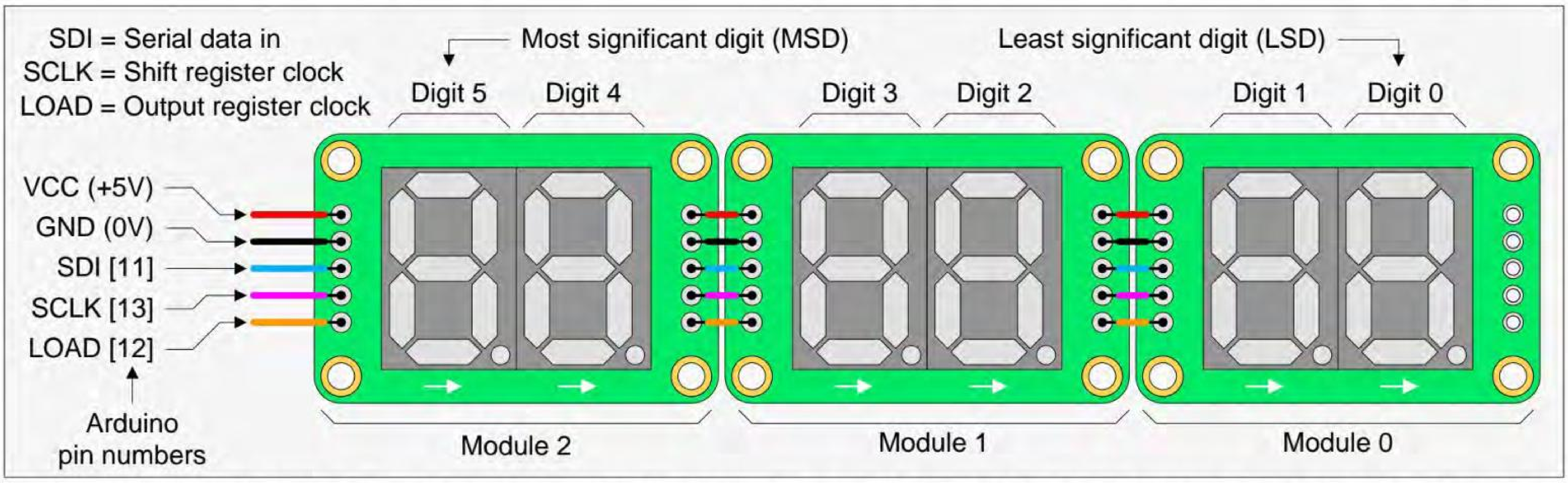

SDI = Serial data in
SCLK = Shift register clock
LOAD = Output register clock
Most significant digit (MSD)
Least significant digit (LSD)
Digit 5
Digit 4
Digit 3
Digit 2
Digit 1
Digit 0
VCC (+5V)
GND (0V)
SDI [11]
SCLK [13]
LOAD [12]
Arduino pin numbers
Module 2
Module 1
Module 0

Fig.2: how I connected the display modules together. Your connections will be similar but it will depend on the exact modules used.

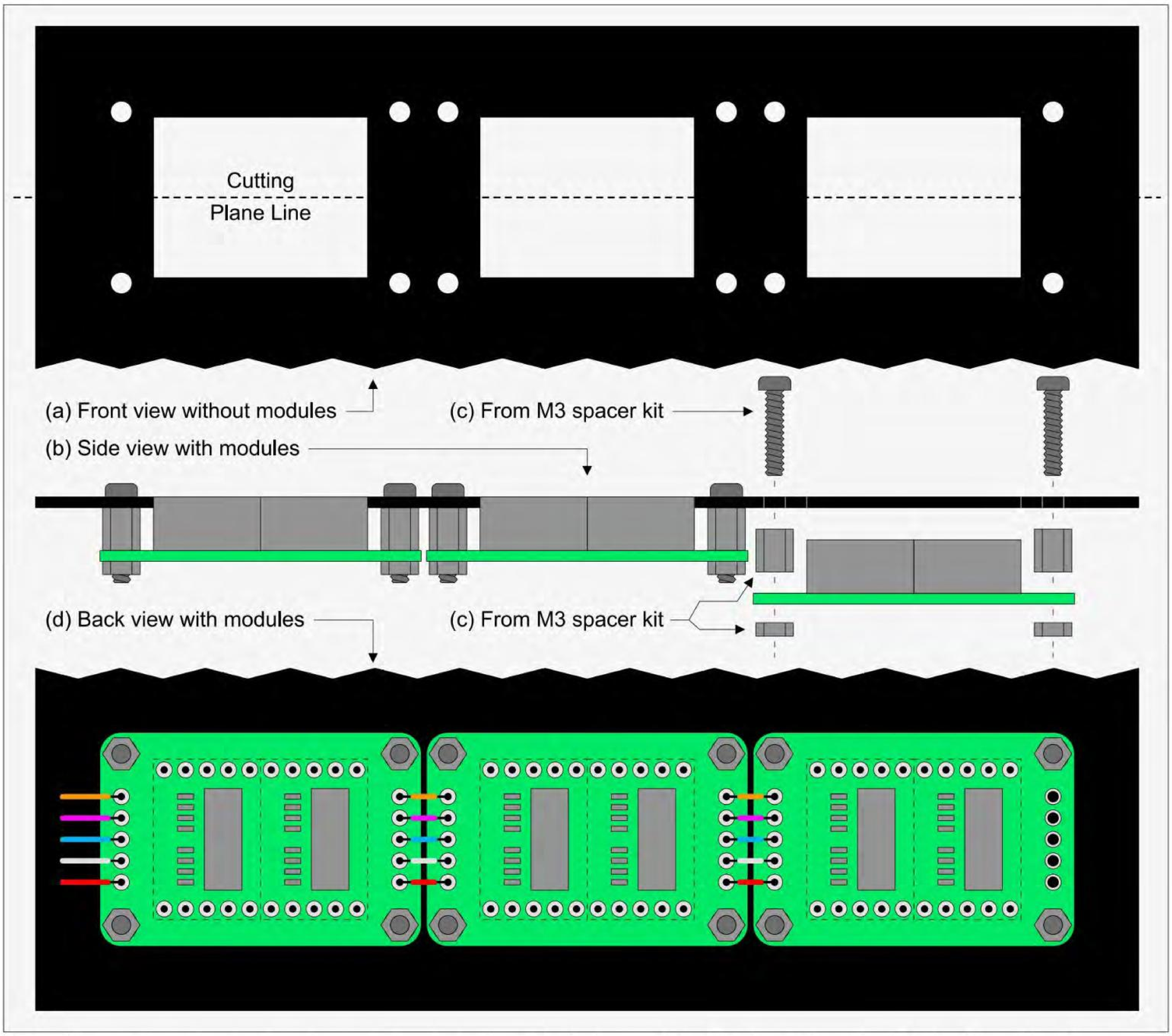

Cutting
Plane Line
(a) Front view without modules
(c) From M3 spacer kit
(b) Side view with modules
(d) Back view with modules
(c) From M3 spacer kit

Fig.3: different views showing how I mounted and wired up the display modules. Don't cut along the "Cutting Plane Line!".

too fast for you to see what's happening on your own setup, you can always increase the **UPDATE_RATE** value to, say, 1000 (one second) or more.

The way to think about this is that the program is physically writing to the two digits on the left of the display (digits 5 and 4).

When we shift a new value into these digits, their original contents are shifted into the two digits in the middle (digits 3 and 2).

Similarly, the original contents of the two digits in the middle are shifted into the two digits on the right (digits 1 and 0).

For my next test (in the file named **CB-jan26-code-02.txt**), I decided to use all six digits. I reduced the value of **UPDATE_RATE** to 10 (ten milliseconds, or a hundredth of a second), I changed **NUM_DIGS** ('number of digits') from 2 to 6, and I changed the **MIN_SCORE** and **MAX_SCORE** values to -99999 to 999999, respectively. I also modified the call to the **NumberToBuffer()** function to read:

```
NumberToBuffer(P1Score, 6, 0);
```

This produces the following sequence:

```
ΔΔ ΔΔ ΔΔ
-9 99 99
-9 99 98
-9 99 97
 :  :  :
-0 00 01
00 00 00
00 00 01
 :  :  :
99 99 98
99 99 99
99 99 98
 :  :  :
```

For my next test (in the file named **CB-jan26-code-03.txt**), I decided to present a 4-digit number in the centre of the display, leaving digits 0 and 5 blank. In this case, I set the **MIN_SCORE** and **MAX_SCORE** values to 0 and 9999, respectively. Also, I modified the call to the **NumberToBuffer()** function to read:

```
NumberToBuffer(P1Score, 4, 1);
```

As you would expect (or hope), this results in the following sequence:

```
ΔΔ ΔΔ ΔΔ
Δ0 00 0Δ
Δ0 00 1Δ
Δ0 00 2Δ
 :  :  :
Δ9 99 8Δ
Δ9 99 9Δ
Δ9 99 8Δ
 :  :  :
```

My next test (in the file named **CB-jan26-code-04.txt**) was a variation on this theme. All I did was change the call to the **NumberToBuffer()** function to:

```
NumberToBuffer(P1Score, 4, 2);
```

This produces the following sequence:

Photo 2: a droolworthy display if ever I saw one.

```
ΔΔ ΔΔ ΔΔ
00 00 ΔΔ
00 01 ΔΔ
00 02 ΔΔ
 :  :  :
99 98 ΔΔ
99 99 ΔΔ
99 98 ΔΔ
 :  :  :
```

For my final experiment (for now), I went for broke (see the file named **CB-jan26-code-05.txt**). In this case, I decided that my virtual Player One was playing a game with nine levels numbered 1 to 9. For each level, the score ranges from 0000 to 9999.

I display the letter 'L' (for level) in the most-significant digit (digit 5) and the level itself in digit 4. Meanwhile, the score is presented on digits 0 through 3. This results in the following sequence:

```
ΔΔ ΔΔ ΔΔ
L1 00 00
L1 00 01
 :  :  :
L1 99 98
L1 99 99
L2 00 00
L2 00 01
 :  :  :
L9 99 98
L9 99 99
L1 00 00
 :  :  :
```

Photo 2 shows this program running on my real-world display.

### A haze of confusion

Before we proceed further, I'm afraid we may need to 'wind things back' just a little. Don't worry; this is going to be interesting. Honest!

I recently received an email from a reader called Walter. He is confused about the difference between the bitwise operators (~ [NOT], & [AND] and | [OR]) and the logical operators (! [NOT], && [AND] and || [OR]).

Walter has been experiencing several 'bad hair days' recently because he's also confused about how we perform both mathematical and bitwise operations on the contents of the same variable.

If poor old Walter is baffled and bewildered, he's probably not alone, so I think it behooves us to take a few moments to ensure we are all tap-dancing to the same skirl of the bagpipes.

Apart from anything else, the stuff I'm about to tell you will make some of the things we've already been doing with respect to bit manipulation and bit shifting operations a lot clearer. All of this will come in handy in the future when we start creating games for our retro gaming console.

### Setting the scene

For the sake of these discussions, let's assume that we've already declared three 8-bit **unsigned** integer variables (of type uint8_t) called a, b and y.

We could use any of the following statements to assign a value to variable a:

```
a = 85;          // Decimal
a = 0x55;        // Hexadecimal
a = 0b01010101;  // Binary
```

In all cases, the eight bits that form variable a will end up containing 01010101. Similarly, we could use any of the following statements to assign a value to variable b:

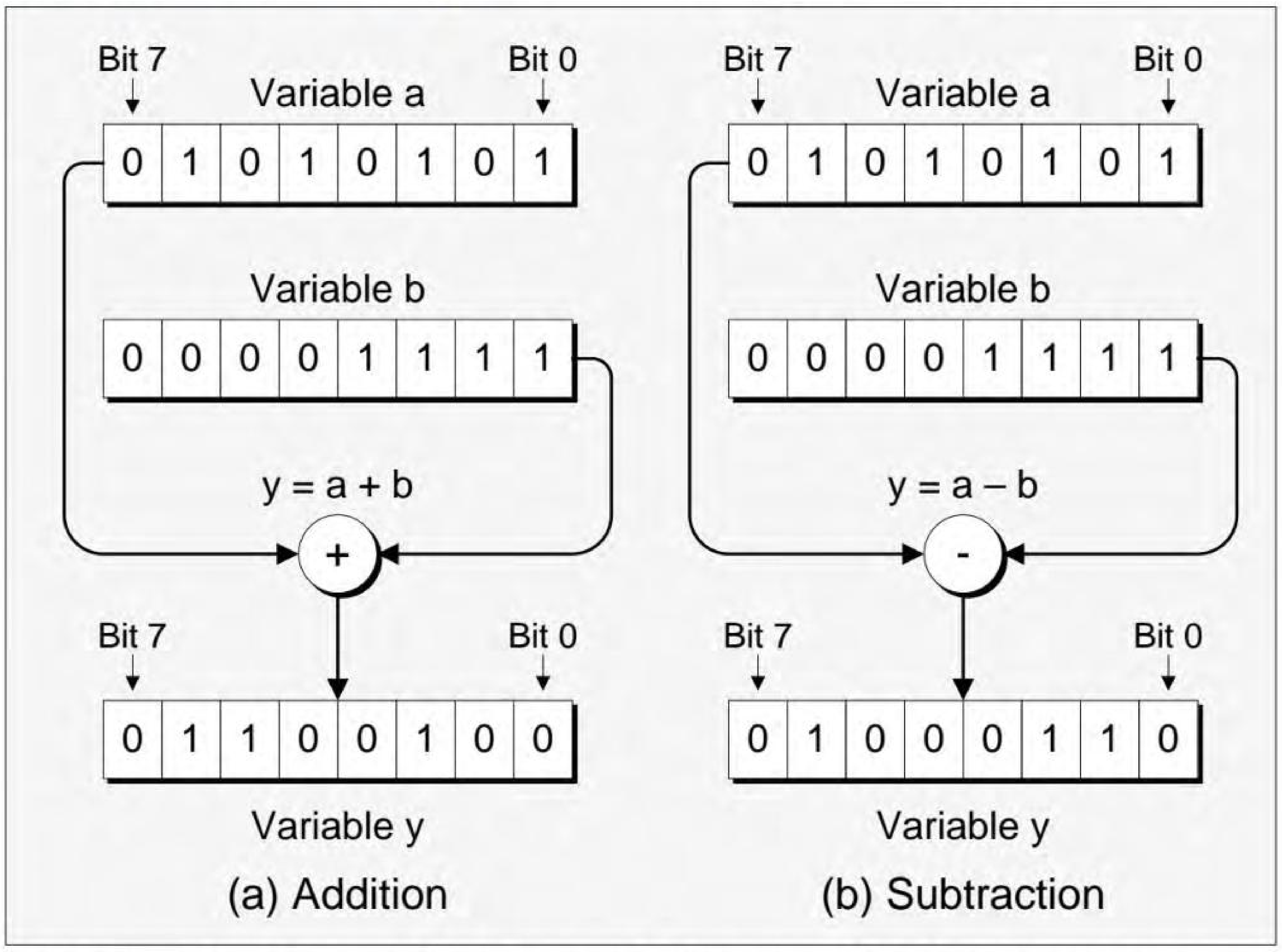
```
b = 15;          // Decimal
b = 0x0F;        // Hexadecimal
b = 0b00001111;  // Binary
```

In all cases, the eight bits forming variable b will end up containing 00001111.

How we view and manipulate these variables is entirely up to us; for example, we can treat them as numbers or as collections of bits. Furthermore, we can switch between these views and mix them as we wish.

### Mathematically speaking

We can perform a variety of mathematical operations on the contents of our integer variables, including + (addition), - (subtraction), * (multiplication), / (division), and % (modulo). For example, consider addition, as shown in Fig.4(a), and subtraction, as depicted in Fig.4(b).



(a) Addition

(b) Subtraction

Fig.4: example calculations of the sums and differences of 8-bit binary numbers.

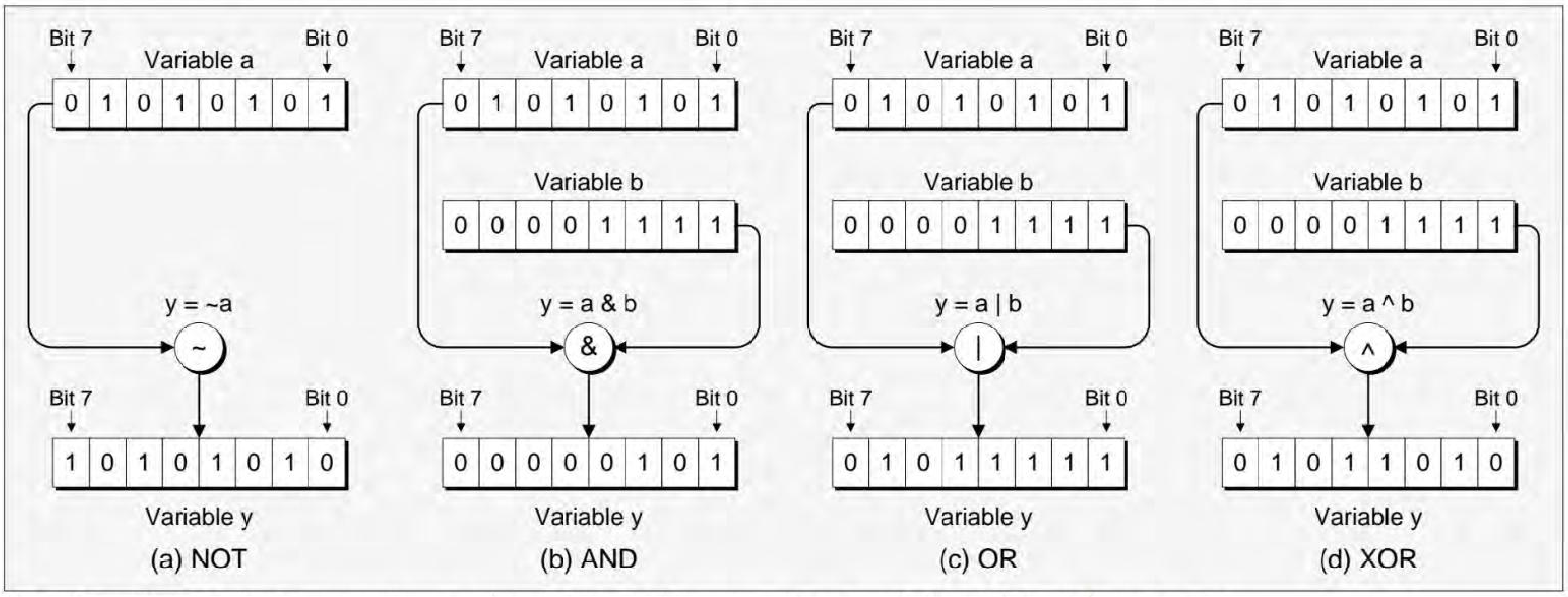


Fig.5: four different bitwise operations performed on 8-bit integer values: not (~), and (&), or (|) and xor (^).

Not surprisingly, 85 + 15 = 100 in decimal (64 in hexadecimal or 01100100 in binary). Similarly, 85 – 15 = 70 in decimal (46 in hexadecimal or 01000110 in binary).

### Feeling a bit bitwise

Once again, let's assume that we start with our a variable containing binary 01010101 and our b variable containing binary 00001111. There are four operators known as the 'bitwise operators': ~ (NOT), & (AND), | (OR), and ^ (XOR). Examples of these operations performed on our 8-bit variables are shown in Figs.5(a), (b), (c), & (d), respectively.

The ~ (NOT) bitwise operator is a unary operator, which means it performs its magic on a single operand (value or variable). As a point of interest, 'unary' comes from the Latin root unus, meaning "one". The other bitwise operators are binary operators, which means they act on two operands ('binary' comes from the Latin bini, meaning "two at a time" or "in pairs").

The word "bitwise" comes from two parts: "bit", short for binary digit, which is the smallest unit of data (0 or 1), and "wise", an English suffix meaning "in the manner of" or "with respect to".

So, "bitwise" literally means "in the manner of bits" or "with respect to individual bits."

Bitwise operators treat their operands as raw bit patterns rather than numerical values. When we perform a bitwise operation, the CPU manipulates each bit independently; it's not working on the number as a single value, but rather bit by bit.

In the case of the & (AND) operation shown in Fig.5(b), for example, bit 0 of a is ANDed with bit 0 of b, and the result is stored in bit 0 of y; bit 1 of a is ANDed with bit 1 of b, and the result is stored in bit 1 of y, and so forth.

All these individual AND operations take place simultaneously, so the operation (a & b) consumes only a single Arduino clock cycle when working with 8-bit values.

### Logically speaking

There are three more operators: ! (NOT), && (AND), and || (OR). These operators evaluate the 'truth values' of their operands (which we can think of as 'inputs') and produce a truth (Boolean, or true/false) value as a result.

In the case of the operands (inputs), a zero (0) value is considered false, while any non-zero value is considered true. By comparison, the output from a logical operator is always 0 (false) or 1 (true).

Let's start with the ! (NOT) operator, depicted in Fig.6. The first case shown in Fig.6(a) isn't surprising; a 0 (false) is transformed into a 1 (true). Similarly, in Fig.6(b), a 1 (true) is transformed into a 0 (false).

It's the example shown in Fig.6(c) that can trip up beginners. In this case, even though bit 0 of a is 0, variable a as a whole contains a non-zero (true) value, which will be transformed by the ! (NOT) operator into a 0 (false) output.

The && (AND) logical operator evaluates the truth values of its operands. If **both** are non-zero, (true), then the result is 1 (true); otherwise, the result is 0 (false). For example, if a = 5 (non-zero, true) AND b = 9 (non-zero, true), then the result of y = a && b will be y = 1 (true). However, if **either** (or both) a or b is 0 (false), the result of y = a && b will be y = 0 (false).

The || (OR) logical operator also evaluates the truth values of its operands. If **either** is non-zero (true), the result is 1 (true); otherwise, the result is 0 (false). For example, if a = 5 (non-zero, true) OR b = 9 (non-zero, true), the result of y = a || b will be y = 1 (true). However, if both a and b are 0 (false), the result of y = a || b will be y = 0 (false).

It's important to note that these operators can also be used to ask questions, like:

```
if (a && b) {}
if (a || b) {}
```

If the result of the logical operation is true (1), then all the statements contained in the squiggly brackets { } will be evaluated; otherwise, the processor will skip over them and continue to the next statement in the program.

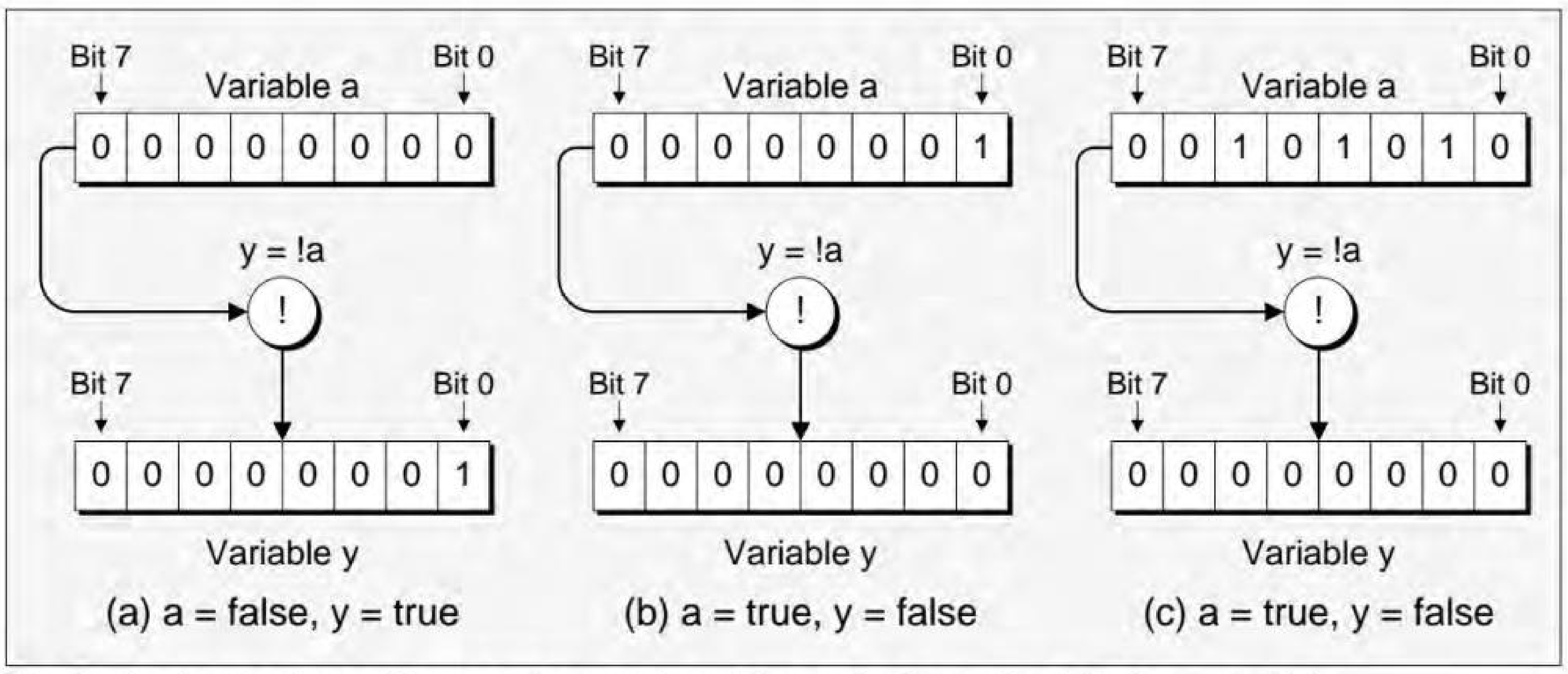
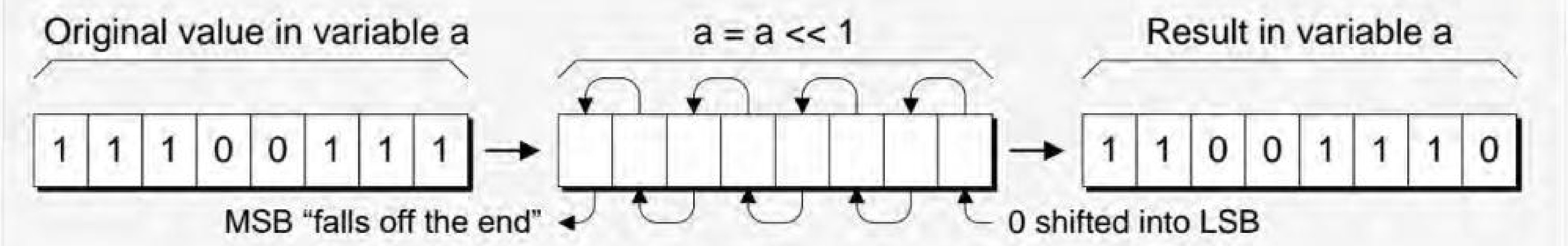
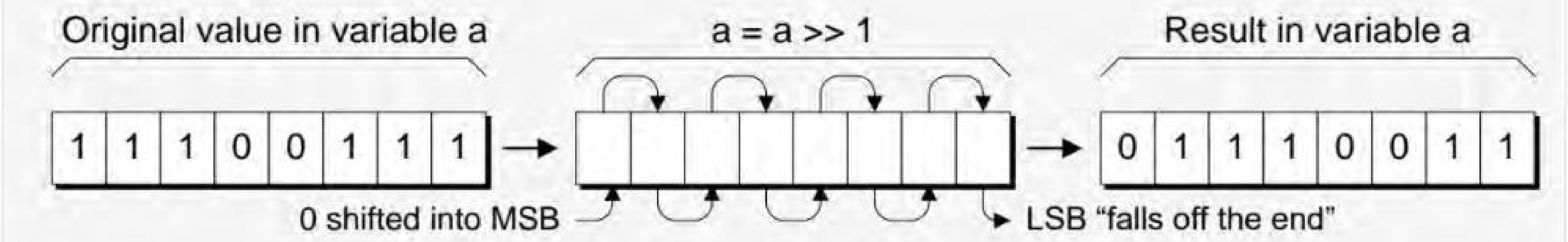


Fig.6: the logical not (!) operator works differently from the bitwise one (~).

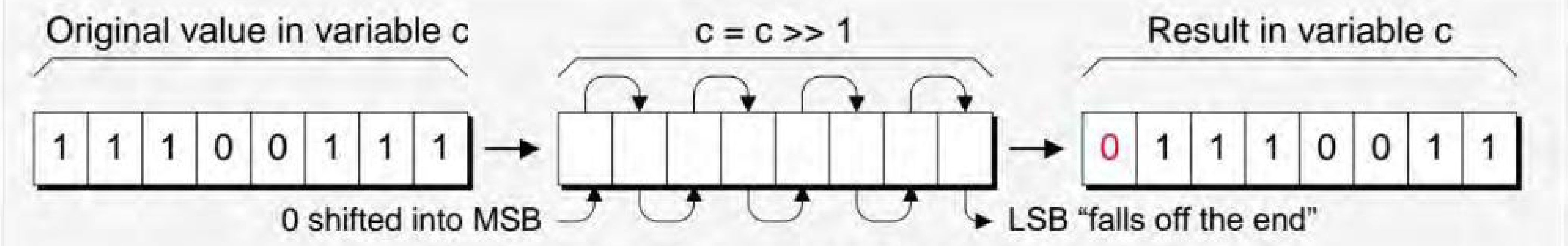


(a) logical shift left on unsigned 8-bit value (uint8_t)

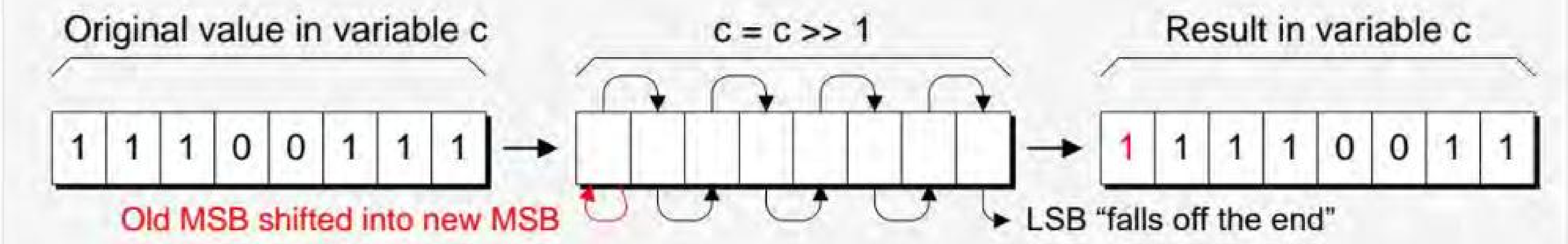


(b) logical shift right on unsigned 8-bit value (uint8_t)

Fig.7: logical bit shifts (left and right) on unsigned 8-bit integers.



(a) logical shift right on signed 8-bit value (int8_t)



(b) arithmetic shift right on signed value (int8_t)

Fig.8: the two possible types of right shift options on signed integers (8-bit in this case).

**Shift over!**

Now, let’s assume that we load our 8-bit unsigned variable `a` with a value of 11100111 in binary (E7 in hexadecimal, 231 in decimal). We can use the `<<` (left shift) or `>>` (right shift) operators to shift this value to the left or right (Fig.7). The shift operators belong to the same general category as the bitwise logical operators on the basis that they manipulate the contents of variables at the bit level.

In this example, we are shifting by only a single bit (eg, `a = a << 1`), but we can shift by as many bits as we wish (eg, `a = a << 3`). Irrespective of the number of bits shifted, these instructions will consume only a single Arduino Uno clock cycle when performed on 8-bit variables.

The shifts performed on unsigned integers are always “logical shifts”, meaning that the values shifted in are always 0s, irrespective of the shift direction. If only life were always this simple, but, of course, it isn’t, as we shall see…

Assume that we declare an 8-bit **signed** integer variable (of type `int8_t`) called `c`. Also, assume that we load this variable with the same binary pattern of 11100111 that we started with before. This still equates to E7 in hexadecimal; however, it now represents –25 in decimal (we’ll discuss the rationale behind the signed binary format further in our next column).

If we perform a **left shift** on a signed integer, then this will be implemented as a logical shift; that is, 0s will be shifted into the least significant digits. However, if we perform a **right shift** on a signed integer, there are two possibilities, as illustrated in Fig.8.

The first possibility is a **logical right shift**, in which 0 is copied into the most significant digit(s), as illustrated in Fig.8(a). This is equivalent to a logical right shift on an unsigned integer. The second possibility is known as an arithmetic shift right. In this case, the original value in the MSB is copied into the most significant digit(s), as illustrated in Fig.8(b).

The reason we might prefer an **arithmetic right shift** over a **logical right shift** with signed integers is too involved to go into here (we’ll discuss this further in our next column). However, this point is largely academic because the silly sausages who created the original C (and later C++) specifications left this choice unspecified, leaving it up to the compiler implementers.

As a result, we, the users, have no clue what to expect (although most modern compilers do the logical thing and use an arithmetic shift in this case; we’ll explain why that makes sense next month). This is why I usually declare my integer variables as unsigned whenever I intend to perform bitwise operations on them.

| Operator | Description | Example |
|---|---|---|
| == | Equal to | a == b |
| != | Not equal to | a != b |
| < | Less than | a < b |
| <= | Less than or equal to | a <= b |
| > | Greater than | a > b |
| >= | Greater than or equal to | b >= b |

Fig.9: C/C++'s main comparison operators.

**Beyond compare**

The C/C++ programming languages provide a suite of equality and relational operators; the main ones are shown in Fig.9. These operators are used to compare two values; for example, to check whether one is greater than, less than, or equal to another. Each comparison produces a Boolean result: `true` (1) or `false` (0).

These operations allow us to ask questions, like the following:

```
if (a == b) {}
if (a != b) {}
```

If the result of the comparison is `true` (1), all the statements in the squiggly brackets { } will be evaluated; otherwise, the processor continues on to the next statement. This also allows us to ask questions like:

```
if (a == true) {}
```

If `a` is a Boolean value (true or false), that is the same as asking:

```
if (a) {}
```

Similarly, we could say:

```
if (a == false) {}
```

or

```
if (a != true) {}
```

Assuming `a` is Boolean, these are the same as asking:

```
if (!a) {}
```

Of course, we can use the logical operators to combine multiple comparisons, such as:

```
If ((a > b) && (c == d)) {}
```

You can read that as “if `a` is greater than `b` and `c` equals `d`, do something; otherwise, don’t”.

**Without precedence**

I mentioned this a couple of years ago, but it’s well worth repeating. In the context of computer programming languages like C and C++, the term

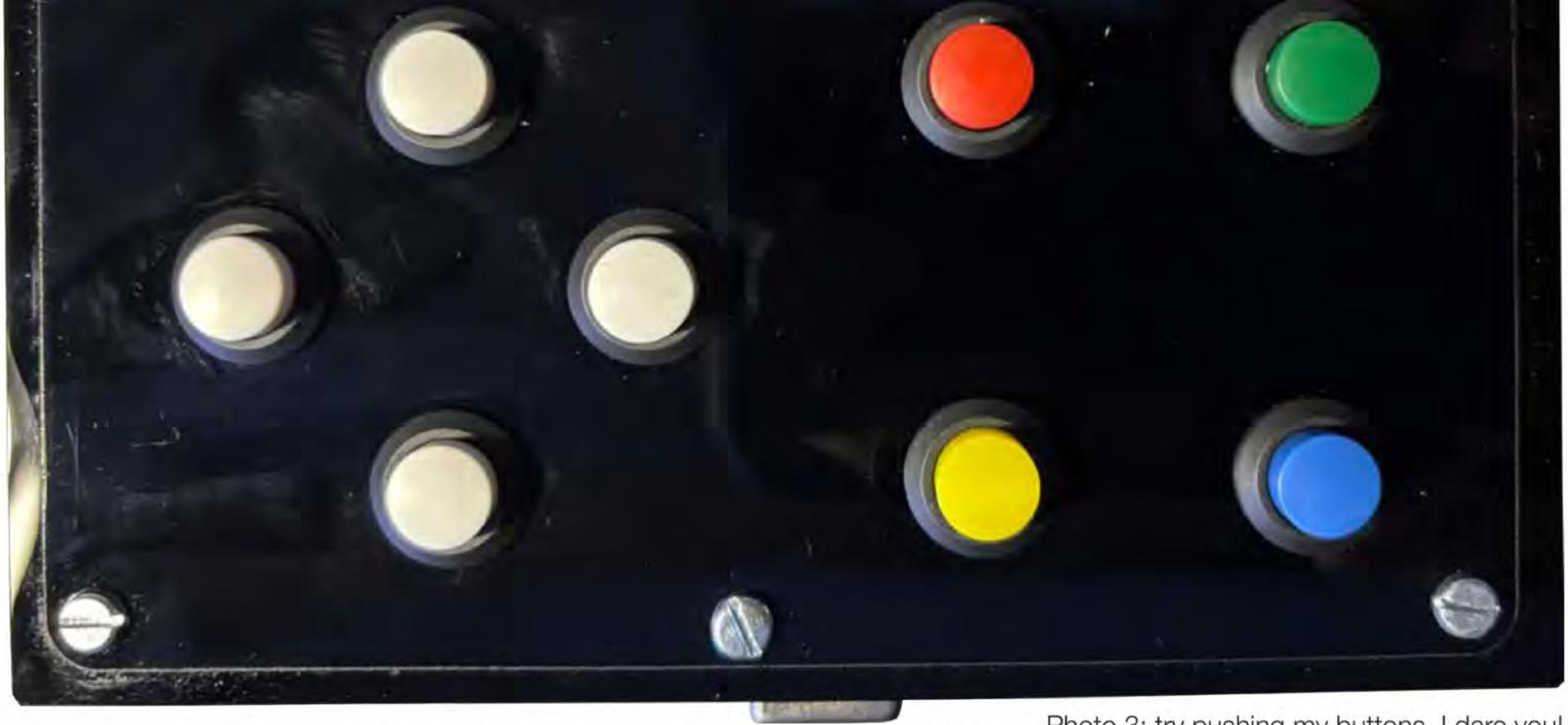


Photo 3: try pushing my buttons, I dare you!

“precedence of operators” refers to the order in which operations are performed.

For example, the multiplication, division, and modulo operators (`*`, `/`, and `%`) have higher precedence than the addition and subtraction operators (`+` and `–`), and all of these arithmetic operators have higher precedence than the logical and relational operators discussed above (well, except for the unary operators `~` and `!`).

You may recognise this as being similar to the basic rules of arithmetic, eg, 2 + 3 × 4 = 14, not 24, because you multiply before adding.

To be honest, this can be a bit of a minefield. For example, `a + b == 7` is evaluated the same as `(a + b) == 7`, while `a & b == 7` is evaluated identically to `a & (b == 7)`.

Now, how would you expect the condition `a + 6 > b / 2 && c * 3 < d – 5` to be evaluated? Suffice it to say that this is why you’ll find my code includes more than its fair share of brackets ( ) to enforce the precedence I require (see https://w.wiki/FvmS for more details).

| Operation | Standard Assignment | Compound Assignment |
|---|---|---|
| Addition | a = a + b; | a += b; |
| Subtraction | a = a - b; | a -= b; |
| Multiplication | a = a * b; | a *= b; |
| Division | a = a / b; | a /= b; |
| Modulo | a = a % b; | a %= b; |
| Bitwise AND | a = a & b; | a &= b; |
| Bitwise OR | a = a \| b; | a \|= b; |
| Bitwise XOR | a = a ^ b; | a ^= b; |
| Shift Left | a = a << b; | a <<= b; |
| Shift Right | a = a >> b; | a >>= b; |

Fig.10: the compound operators in C/C++.

### Compounding the problem

Since we’ve already covered such a lot of ground, we may as well remind ourselves that C and C++ provide a set of compound operators that offer a shorthand way of combining an operation with an assignment (see Fig.10).

Observe that there are no `~=` or `!=` operators listed in that table because bitwise NOT (`~`) and logical NOT (`!`) are unary, not binary—ie, they act on a single operand (although you can write lines like `a = ~a;` or `a = !a;`).

Also, for reasons that are obvious when we think about it a little, there are no compound assignment forms for the logical AND (`&&`) and logical OR (`||`) operations, so we always need to write these out explicitly.

### Devices and components used in this issue

| | |
|---|---|
| **Arduino Uno R3 microcontroller module** | https://pemag.au/link/ac2g |
| **Solderless breadboard** | https://amzn.to/3O2L3e8 |
| **4-inch (10cm) jumper wires (optional)** | https://pemag.au/link/ac2l |
| **8-inch (20cm) jumper wires (male-to-male)** | https://amzn.to/3O4hnxk |
| **LEDs (assorted colours)** | https://amzn.to/3E7VAQE |
| **Resistors (assorted values)** | https://amzn.to/3O4RvBt |
| **Ceramic capacitors (assorted values)** | https://pemag.au/link/ac2i |
| **16V 100µF electrolytic capacitors** | https://pemag.au/link/ac2j |
| **8-Segment DIP red LED bar displays** | https://pemag.au/link/ac2c |
| **74HC595 8-bit shift registers** | https://pemag.au/link/ac1n |
| **Dual 7-segment display modules** | https://pemag.au/link/ac8i |
| **0.1” pitch header pins** | https://pemag.au/link/ac8j |
| **PCB spacer kit (black, M3)** | https://pemag.au/link/ac4u |

### Educational? I should say so!

It probably goes without saying (but I’ll say it anyway) that one of the advantages of subscribing to a magazine like *Practical Electronics* is that you get material like the above. This is like a canned and distilled version of an educational course that you’d normally pay good money for. It’s just one more service I provide. You’re welcome.

### Next time

I ended my last column saying, “... despite previous broken promises, we really will add the eight push-button switches to our console’s front panel”.

Although I didn’t actually find the time to wire these little rascals up, I have indeed attached them to my front panel, as shown in Photo 3, which means I can’t be accused of “prevaricating around the bush”, as Wallace might say. Still, more on that soon!

As always, if you have any thoughts you’d care to share on anything you’ve read here, please feel free to email me at max@clivemaxfield.com

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### ~ Tim Blythman's NFC Programmable ~

# IR Remote Control Keyfob

**Sometimes you need a small infrared (IR) remote control for just a handful of functions. This remote is about the smallest we've seen, it can hang on your keychain and you can make it yourself. It has three buttons that can trigger separate functions that are programmable wirelessly via NFC.**

**We've used universal infrared remote controls** in several projects over the last few years, most recently in the Eight Channel Learning IR Remote Receiver from the August 2025 issue. Another project that used a similar remote control is the Multi-Channel Volume Control from the November and December 2024 issues.

The problem with these is that the remotes tend to be rather large, as they are intended to control devices like TVs and Blu-ray players that have many functions. For these projects, we need just a few functions like volume up, volume down or on/off.

We found a smaller remote from a local retailer that seemed rather useful, but it was discontinued not long after we found it. So we thought it'd be a good idea to design a small IR remote control with just a few buttons for those situations where that's all you need. Rather than having buttons that send fixed IR codes, and rely on the receiver to be able to adapt to that, we felt we could improve it. It would be handy for such a device to be programmable.

The difficulty lies in adding a way to allow codes to be added or changed easily. We don't want to massively complicate the device with a screen, more buttons etc! Nor would it be ideal to build external hardware to plug into a socket on the keyfob. Luckily, there is a neat solution.

You might recall our Dynamic NFC/RFID Tag from July 2024. It combined a small chip with a PCB trace antenna to create a programmable NFC/RFID tag that can be used to hold and transfer small amounts of information.

NFC (near-field communications) is a protocol based on RFID (radio frequency identification) technology. It allows communication with devices over short distances, typically up to 5cm. It's the technology that's used in things like contactless credit cards and transit passes.

Here, we have used NFC to add a programming interface to the IR Remote Control. An external device such as a mobile phone becomes the programmer, and the setting can be transferred wirelessly to the Remote Control, without needing a socket or opening the case!

The Remote Control is simply placed against the NFC reader on a device (eg, on the back of a mobile phone) and an app is used to control the transfer of data. The NFC chip we are using doesn't even need external power, so the data can be transferred without a battery fitted to the Remote Control.

## Other features

Since we are using the same chip as the earlier Dynamic NFC/RFID Tag, you can use this device similarly if you wish. The ST25DV04 has 512 bytes of EEPROM that can be used to store all manner of information, as well as the configuration for the Remote Control. Compatible chips with more storage are also available (up to at least 8kiB).

The NFC protocol allows up to four different NDEF (NFC Data Exchange Format) records to be stored. Programming the Remote Control only requires a single text format NDEF record to hold the programming data; the remaining space can be used to store any other information you want.

For example, an NDEF record can contain a URI to link to a webpage, or a WiFi record that contains the information needed to connect to a WiFi network. It could even contain a virtual business card, embedding data relating to contact details and phone numbers.

The MIME record type could contain a complete file, such as an image, although its utility is somewhat limited by the small amount of memory on the chip.

So you could also use the RFID Programmable IR Fob Remote as a portable NFC tag which can be used to pass around information such as webpage links or virtual business cards. The data is transferred by simply tapping the fob against an NFC reader.

## Circuit details

Fig.1 is the complete circuit diagram. Power comes from coin cell BAT1 in a holder, which has a 22µF capacitor across it. This relatively large capacitance helps to even out the demands on the coin cell. Its lifespan can be adversely affected by high loads.

A further 100nF capacitor provides local bypassing for IC2, a PIC16F15224 microcontroller. This is a fairly basic 8-bit 14-pin part, but it has PWM and timer features to allow the modulation and timing needed to implement an IR transmitter.

This chip also has a very low-power sleep mode, which is handy for a device powered by a small cell. IC2 connects to the CON1 header for in-circuit serial programming (ICSP) at its pins 1, 4, 12, 13 and 14. A 10kΩ pullup on pin 4 sets the microcontroller to run normally unless a programmer overrides this signal.

IC1 is the ST25DV04 dynamic NFC tag chip. Pins 2 and 3 connect to a PCB trace inductor which has a nominal inductance around 4.7µH. When

# IR Remote Control NFC Keyfob

- Compact keyfob case: 61 × 36.5 × 15.8mm
- Can attach to a keyring
- Three buttons to trigger the IR emitter
- Power supply: CR2032 lithium coin cell
- Standby current: <1µA
- Active current: 3mA
- Status indicator: red/green LED
- IR protocols supported: NEC, Sony SIRC and Philips RC5 & RC6
- Low battery indicator
- Integrated NFC tag
- Programmable with ST25 NFC Tap mobile app
- Can work with our Multi-channel Volume Control, Eight Channel Learning IR Remote Receiver and other projects

VOL+
VOL–
MUTE
NFC
NFC PROGRAMMABLE
IR FOB REMOTE λ→
S1
S2
S3
Q1
IC1
IC2
CON1
LED2
22µF
100Ω
100nF
4.7kΩ
2kΩ

combined with the chip's internal 28.5pF capacitance, it is resonant at NFC's 13.56MHz frequency. The trace inductor consists of eight loops on the back of the PCB.

IC1's pin 4 is ground, while power on pin 8 is supplied by IC2's pin 8, along with a 100nF bypass capacitor. This means that IC2 can completely power off IC1 by setting that pin low (although IC1 can still get power from its antenna in that case). Pins 6 and 7 of IC2 connect to pins 5 and 6 of IC1 for the I²C interface; each also has the requisite 4.7kΩ pullup to the switched power line.

Tactile pushbuttons S1, S2 and S3 connect to pins 5, 3 and 2 of IC2 respectively, with the other sides connected to ground. The microcontroller applies an internal pullup to those pins so that the switch state can be detected; these pins are in a high state until the button is pressed, then it goes low.

The remaining circuitry drives the infrared (IR) transmitter LED2, and a bicolour indicator LED1. LED1's red and green junctions are in inverse parallel with other, with both in series with a 2.2kΩ resistor connected between pins 9 and 10 of IC2. By driving one pin high and the other low, either the red or green LED can be lit.

The section around IR LED LED2 has been designed to provide high bursts of current to drive the transmitter, while at the same time enforcing a low average current draw on the coin cell. The 470Ω resistor and 22µF capacitor provide a local buffer, while the 100Ω resistor limits the peak current.

With the values used, the average current draw of LED2 during transmission is 2mA, while the IR LED sees peaks of 15mA, which gives a good compromise between transmission power and the draw on the coin cell. LED1 and IC2 will also draw current while the transmitter is active, adding to the load on the cell.

This simple circuitry only switches the IR LED to be on when IC2's pin 11 takes Mosfet Q1's gate high, and off when pin 11 is low. The microcontroller must modulate the signal to suit the receiver detection frequency and the expected protocol.

## Firmware

The microcontroller runs with a 2MHz instruction clock, much lower than its 8MHz maximum. That reduces its current draw when it is active by about 2mA. If the clock was much slower, the micro would struggle to generate the necessary waveforms for IR transmission.

The micro is normally in sleep mode and it draws less than 1µA. Our Coin Cell Emulator from November 2024 gives a reading of 0.0µA in this state! Thus, the cell life will be dominated by how much the Remote Control is used and the cell's shelf life.

When a button is pressed, the micro 'wakes up' from sleep mode and acts upon the button presses. When the buttons are released and transmission has ceased, the micro checks the supply voltage. If it is 2.6V or higher, the green LED in LED1 is flashed briefly; otherwise, it flashes red. This is a simple but effective battery status indicator.

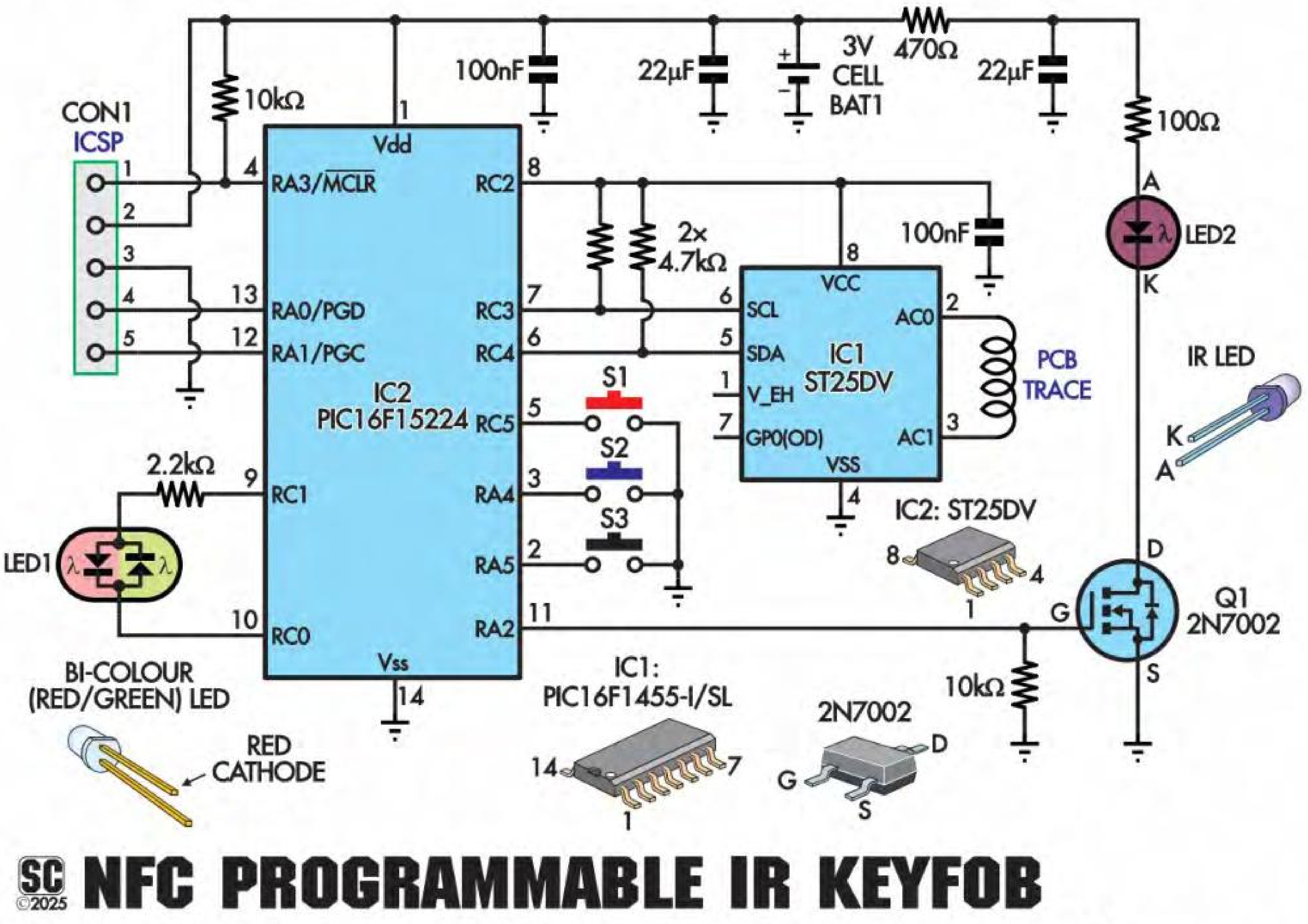

CON1
ICSP
1
2
3
4
5
10kΩ
4
RA3/MCLR
1
Vdd
13
RA0/PGD
12
RA1/PGC
IC2
PIC16F15224
9
RC1
10
RC0
Vss
14
RC2
8
RC3
7
RC4
6
RC5
5
RA4
3
RA5
2
RA2
11
100nF
22µF
3V CELL BAT1
470Ω
22µF
100Ω
A
LED2
K
2× 4.7kΩ
100nF
SCL
SDA
V_EH
GP0(OD)
VCC
AC0
IC1 ST25DV
AC1
VSS
PCB TRACE
S1
S2
S3
2.2kΩ
LED1
λ
BI-COLOUR (RED/GREEN) LED
RED CATHODE
IC1: PIC16F1455-I/SL
14
7
1
IC2: ST25DV
8
4
1
2N7002
D
G
S
10kΩ
Q1 2N7002
IR LED
K
A
SC ©2025
NFC PROGRAMMABLE IR KEYFOB

**Fig.1: the Remote Control circuit is straightforward. IC1 is powered from one of IC2's I/O pins, allowing it to be fully powered off to minimise battery drain. The circuit around Q1 and LED2 allows LED2 to be driven at 15mA peak while limiting the draw on the coin cell to only 3mA.**

The IR transmitter combines a timer and PWM peripheral to generate the IR modulation, which can vary between 36kHz and 40kHz, depending on the protocol. The processor encodes the IR signal as a series of active and inactive phases at the carrier frequency.

The Remote Control supports the NEC, Sony (SIRC) and Philips (RC5 and RC6) protocols. John Clarke explained all of these in detail in the article on the Eight Channel Learning IR Remote Receiver.

An interrupt is triggered on each PWM cycle, providing the timing to step through the active and inactive phases of the encoded signal. LED1 is driven in time with the active phases of the IR signal, to give confirmation that transmission is occurring. The colour reflects the battery state; it flashes green if the battery is fine or red if its voltage is low.

When the transmission cycle ends, the buttons are checked and the IR transmission continues if the button is held down. For most protocols, that means simply repeating the previously sent sequence, but the NEC protocol uses a special repeat packet instead.

## NFC programming

That IC1 can be programmed via its RF interface is completely transparent to the rest of the circuit. Its electrical interface is much the same as many I²C EEPROM devices, although its contents also include a header identifying the size and nature of the data, which needs to be read and validated before the data is processed.

In the event that all three buttons are pressed at the same time, the micro quickly alternates the green and red LEDs to alert the user. When the buttons are released, it powers on IC1 and attempts to read an NDEF text record from its internal memory, then powers down IC1 immediately.

If the read is successful and correctly formatted data is found, the codes are loaded into memory and are available for use the next time any of the buttons are pressed. During this sequence, LED1 blinks in various patterns to report on the status of the programming.

We'll discuss the text format, programming and LED colour codes in more detail later, as well as the use of the ST25 NFC Tap mobile app.

## Construction

The PP43 fob enclosure that we are using for this project comes equipped with buttons (to actuate the switches) and a 3mm hole that suits LED1. However, since these cases are designed to house RF transmitters, they lack a hole for the IR LED.

We recommend adding this hole as the first step, since it will be easier and neater to tweak the location of the IR LED than to modify the hole in the case. You can use the PCB to mark it out, or use the measurements in the Fig.2 drilling diagram.

Firmly tape the two case halves together. The hole is centred on the division between the two halves. We found it worked well enough to drill both halves at the same time as long as they were securely held together. Check our photos to confirm the placement of the hole and start with a smaller pilot hole to locate it accurately.

You can see that the hole sits underneath the black button. It is a 5mm diameter hole and, of course, it goes in the opposite end of the case from where a keyring would attach. The metal battery tabs included with the case are not needed since we are using a cell holder fitted to the PCB.

## PCB assembly

The remaining parts can now be fitted to the PCB. They are mostly SOIC or M3216/1206-sized SMD devices, along with some through-hole parts. So it is easy enough to construct even if you have had minimal experience working with SMD parts.

At a minimum, you will need a syringe of flux paste and tweezers if you are accustomed to working with through-hole parts. Your flux will likely recommend a cleaning solvent; if not, isopropyl alcohol works well for most fluxes (you can use methylated spirits in a pinch).

We also recommend you keep on hand some solder-wicking braid, a magnifier and a fume extraction fan. If you don't have such a fan, work near an open window or outdoors. Working on a uniform light-coloured background will help you find any parts

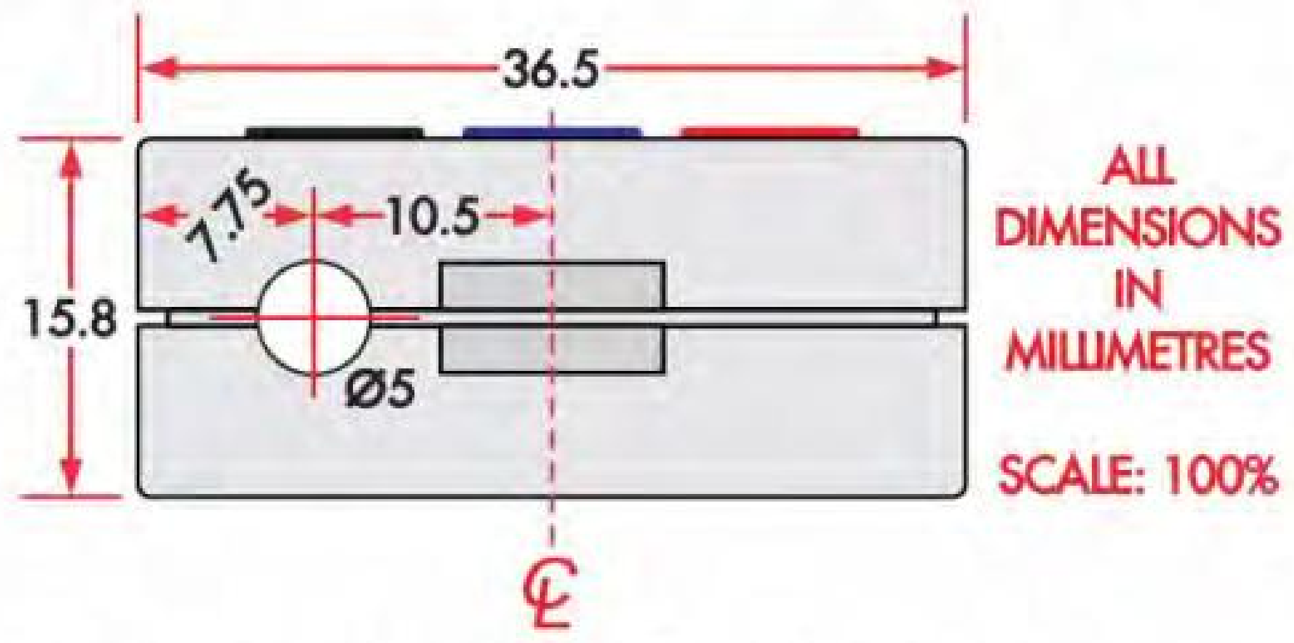
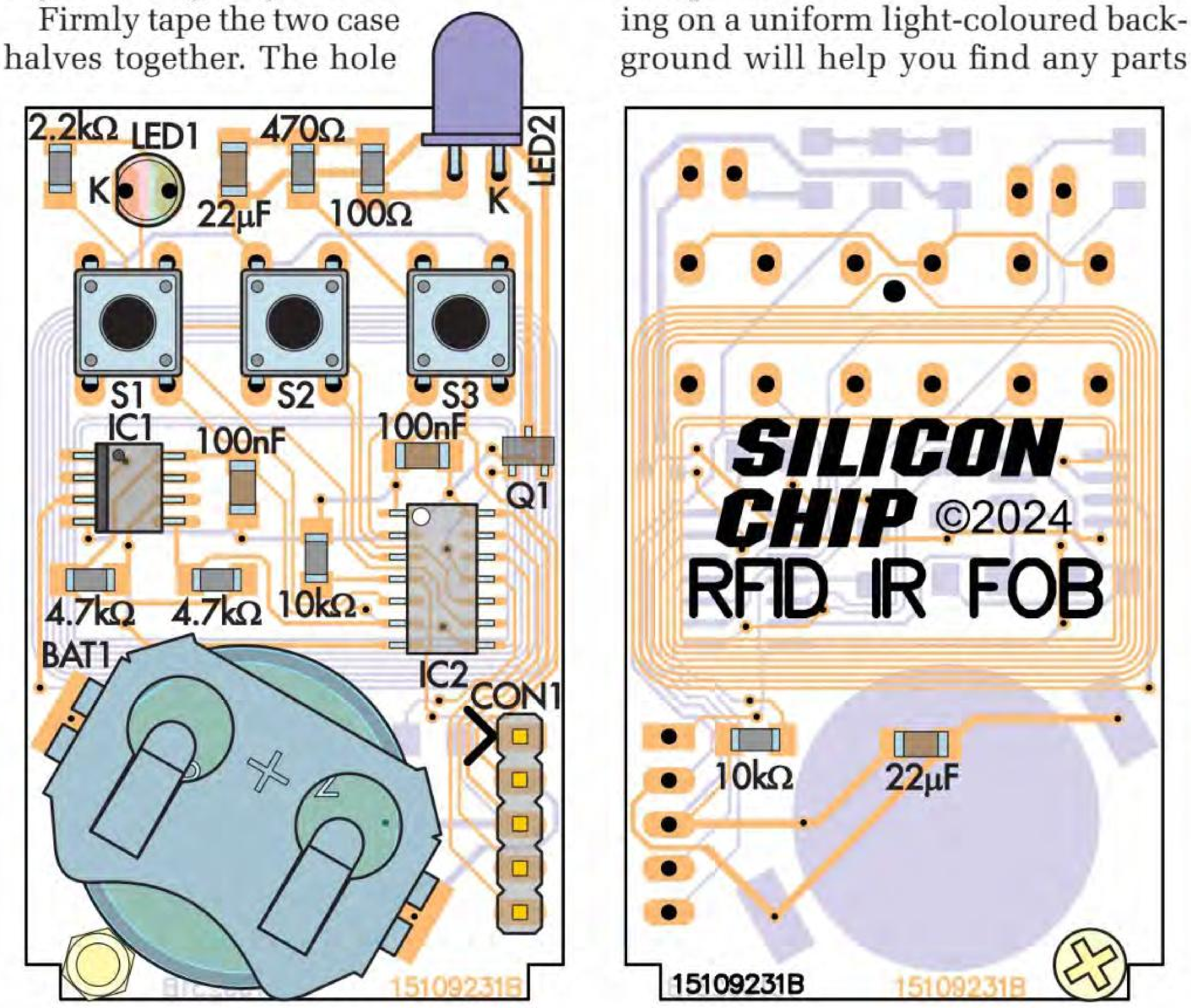


**Fig.2: you can use the dimensions here to locate the hole for the 5mm IR LED, or simply slot the PCB into place and use it to place marks on the case. There is a locating pin on the case to ensure that the PCB is aligned.**



**Figs.3 & 4: most of the components on the front of the PCB are M3216/1206 passives or SOIC chips, so they should be easy enough to solder, even for those inexperienced with surface-mount work. Don't forget to fit the two components on the back of the PCB. The 22µF capacitor helps protect the coin cell from brief high-current demands that could shorten its life. *This diagram is shown at 175% scale.***

that you drop, and a magnifying lamp can also be helpful.

Proceed to fit the parts in the locations shown in the Fig.3 and Fig.4 overlay diagrams. Start with Q1, the only SOT-23 device. Smear a tiny bit of flux paste on its pads (that will make soldering much easier) and rest it in place according to the silkscreen marking, with its leads flat against the PCB.

Tack one lead, then check the positioning of the remaining pins over their pads and adjust as needed by remelting the solder. With it correctly located, solder the remaining leads and refresh the first lead by adding a tiny amount of extra flux paste before touching the iron to it.

The same process can be used to solder IC1 & IC2. Before soldering any pins, it's most important that you identify pin 1 on the IC, which is usually indicated by a dot or divot in one corner. Failing that, look for a chamfered edge along one side. With that side on the left and the writing facing you, pin 1 will be at upper left. Match each chip's pin 1 with the markings on the PCB and Figs.3 & 4.

Solder these parts in place, just like Q1. If you get a solder bridge between two pins, leave it in until all the leads are soldered. To remove a bridge, add more flux, then gently push the braid against the solder with the iron. When it draws in the excess solder, gently slide both away from the part.

Now you can fit the four capacitors using the same technique. There are two 100nF parts, one adjacent to each IC. These will be thinner than the 22µF parts. One 22µF part is near the top of the PCB, while the other is on the back.

The resistors will be marked with value codes (see the parts list). Make sure that the values match the silkscreen and overlay in Fig.3. There are seven resistors to be fitted.

Next, solder the cell holder in place. Align it with the markings on the PCB, being sure that the cell entry faces the corner of the PCB near the BAT1 marker on the silkscreen. You can also compare its orientation to the photos. Take care to line it up correctly, since it may prevent the screw being fitted if it is too close to the edge of the board.

Apply a little solder to one pad, then check its position. If you're happy with that, apply a generous amount of solder to both pads to give mechanical strength.

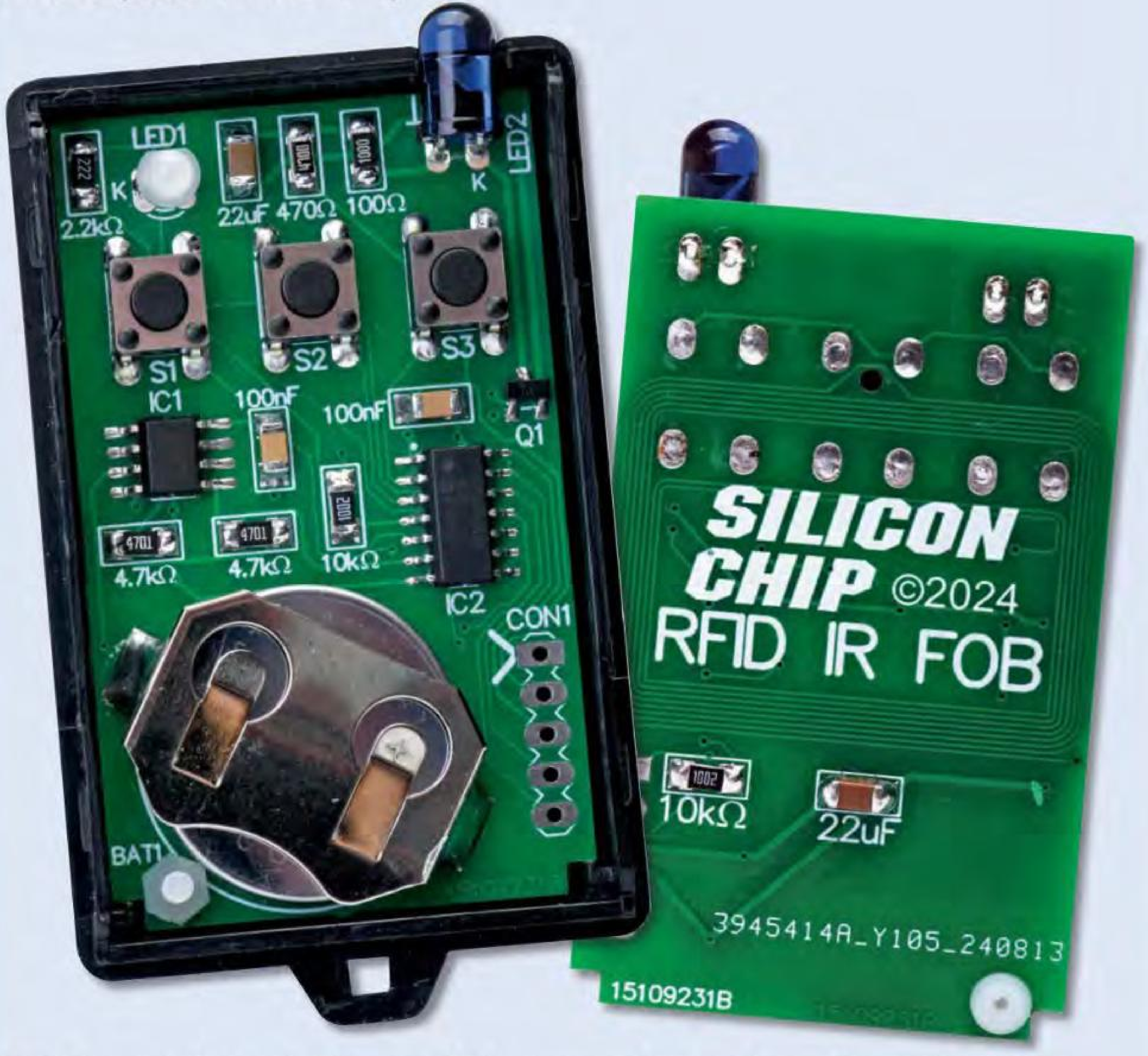
## Parts List – IR Remote Control Keyfob

1 double-sided PCB coded 15109231, 30.5 × 52mm
1 Supertronic PP43 keyfob enclosure
1 2032-size SMD coin cell holder (BAT1) [eg, Linx BAT-HLD-001]
1 CR2032 or CR2025 3V lithium coin cell (BAT1) ♦
1 5-way pin header, 2.54mm pitch (CON1; **optional**, for ICSP) ♦
3 4.3mm-high 6×6mm through-hole tactile switches (S1-S3)
1 M2×6mm Nylon machine screw and hex nut
1 lid label sticker

**Semiconductors**
1 ST25DV04K-IER6S3 (or equivalent) dynamic RFID tag chip, SOIC-8 (IC1)
1 PIC16F15224-I/SL 8-bit microcontroller programmed with 1510923A.HEX, SOIC-14 (IC2)
1 3mm bicolour (red/green) LED (LED1)
1 5mm IR emitter LED (LED2) **[TSAL6200 recommended]**
1 2N7002 N-channel Mosfet, SOT-23 (Q1)

**Capacitors** *(all SMD M3216/1206, X7R ceramic)*
2 22µF 10V
2 100nF 50V

**Resistors** *(all SMD M3216/1206, 1%, ⅛W)*
2 10kΩ (code 1002 or 103)
2 4.7kΩ (code 4701 or 472)
1 2.2kΩ (code 2201 or 222)
1 470Ω (code 470R or 471)
1 100Ω (code 100R or 101)



**This Remote Control will easily fit in your pocket and can trigger up to three different functions. It makes the perfect compact companion for devices like the Eight Channel Learning IR Remote Receiver. The PCB slots into the case and aligns with a pin moulded into its base. The ICSP (in-circuit serial programming) header can be left in place. Note the screw to prevent the coin cell being removed by children. Both images are shown enlarged for clarity.**

***SC7421 Kit (~£15 + P&P):*** *includes all parts listed except the two marked with ♦. The microcontroller is pre-programmed, but the NFC chip will be blank.*

That completes the surface-mounting parts, so clean off any excess flux using your solvent and allow the board to dry thoroughly. Take the time to inspect it under magnification for bridges or bad solder joints, since they will be easier to correct now than later.

## Through-hole parts

Bend IR LED2's leads at right angles directly behind the body. Make sure they are bent in the right direction, such that the shorter cathode lead will go into the hole marked K on the PCB. Push it into the holes and solder one of the leads so that the lens points out parallel to the PCB, then trim both leads (leave the unsoldered one long enough to solder later).

**The hole shown in Fig.2 allows the IR emitter LED to poke out through the end of the case, as seen here. Note the location relative to the buttons.**

You can now place the PCB in the bottom of the case and confirm that LED2 lines up with the hole. Having only one lead soldered will make it easier to adjust the position. Once it is aligned with the hole, solder the other lead.

Next, fit the three tactile switches. These must be no more than 4.3mm high; any taller and they would be permanently depressed by the case buttons. They must also be mounted flat against the PCB for the same reason.

You can now use the PCB along with the top half of the case to check the position of the bicolour LED, LED1. Its top lip should sit no more than 7mm above the PCB. The K cathode marking on the PCB corresponds to the cathode of the red LED in the package.

Test this with a multimeter set to diode mode. When the LED lights up red, the pin connected to the multimeter's black lead is the one that should be placed in the hole marked K.

If you have a pre-programmed chip for IC2, you can fit a coin cell and test the LED's operation. Pressing and holding one of the buttons should cause the LED to flicker and flash green, assuming a fresh cell has been fitted. If it's red instead, you should swap its leads.

## Programming IC2

The ICSP header is only needed if you have to program a blank chip for IC2. A standard height (11mm total) header strip will not foul the case when fitted, so we recommend that you solder this in place and leave it; it will not affect the operation at all.

The software we use for programming PICs is Microchip's IPE (integrated programming interface), which can be downloaded as part of the MPLAB X IDE from www.microchip.com/en-us/tools-resources/develop/mplab-x-ide

A Snap, PICkit 4 or PICkit 5 programmer can be used for programming. A coin cell should be fitted to provide power if needed; for example, the Snap cannot provide power. Connect the programmer to the CON1 header, aligning the pin 1 arrows on the programmer and PCB.

Choose the PIC16F15224 as the part, click connect and confirm that communication is established. If powering IC2 from the programmer (PICkit), you will need to enable that before clicking the connect button. Load the HEX file, program it into the chip and check that it verifies correctly. The LED should briefly flash green as programming finishes or when you disconnect the programmer.

## Final assembly

Fit the cell (+ side up) and secure it with the machine screw and nut through the adjacent hole, feeding the screw from the bottom of the PCB. The case should neatly snap together around the PCB. Check that the buttons actuate correctly and the LED lights up as described earlier.

If the LED lights when no button is pressed, one of the switches may be stuck. Check the solder joints on the back of the PCB and trim down any that are too tall. The PCB should slot neatly into the case and sit flat.

The default programming is to suit the Multi-channel Volume Control, with the red and blue buttons increasing and decreasing the volume, respectively (think hotter and colder!). The black button controls the mute function. The default codes are for an NEC device at address 0, with command codes 21 (red), 7 (blue) and 67 (black).

You might see these values reported differently on some systems. An example is the Micromite or PicoMite IR decoder, which will report codes 168, 224 and 194 respectively because it uses a reversed bit order. The device code is still reported as 0 as the bits are the same when reversed.

Simple hardware, such as the IR Keyboard we created in August 2019 ("Control your PC with an infrared remote") can also be used to interact with this and other IR transmitters. The excellent irremote Arduino library makes it easy to receive all sorts of IR signals.

## Programming the NFC chip

To use the Remote Control with other hardware, you will need to program it to use new codes. First, you need to determine the protocols and codes to use. If you do not have a manual or other reference for these, hardware similar to the IR Keyboard can be used to read codes from an existing remote.

The Arduino irremote library comes with a sample sketch called "ReceiveDump", which reports the protocol and details of received IR signals. We used this extensively during our testing of the Remote Control to check that it was delivering the correct codes.

The NDEF text record required to program the Remote is much the same as a CSV (comma separated variable) file. The first field in each row is a code that identifies the protocol; the codes and protocols are listed in Table 1. The next field is the address code in decimal, followed by the command or data field, also in decimal.

You will need the ST25 NFC Tap mobile app and a device that has NFC capabilities. We used an Android phone and downloaded the app from the Play Store (https://pemag.au/link/ac38). We haven't tested it, but the app also appears on the Apple App Store (https://pemag.au/link/ac39).

There may be other apps that will work; we previously tried the NXP TagInfo and TagWriter apps. Any app that can read and write NFC NDEF records should work.

Screen 1 shows the welcome screen for the ST25 NFC Tap. Hold the back of the Remote Control against the back of the phone (or other device). Screen 2 shows what you will see when the NFC tag in the Remote Control is read.

**Table 1 – protocol codes for RFID Programmable IR Fob Remote**

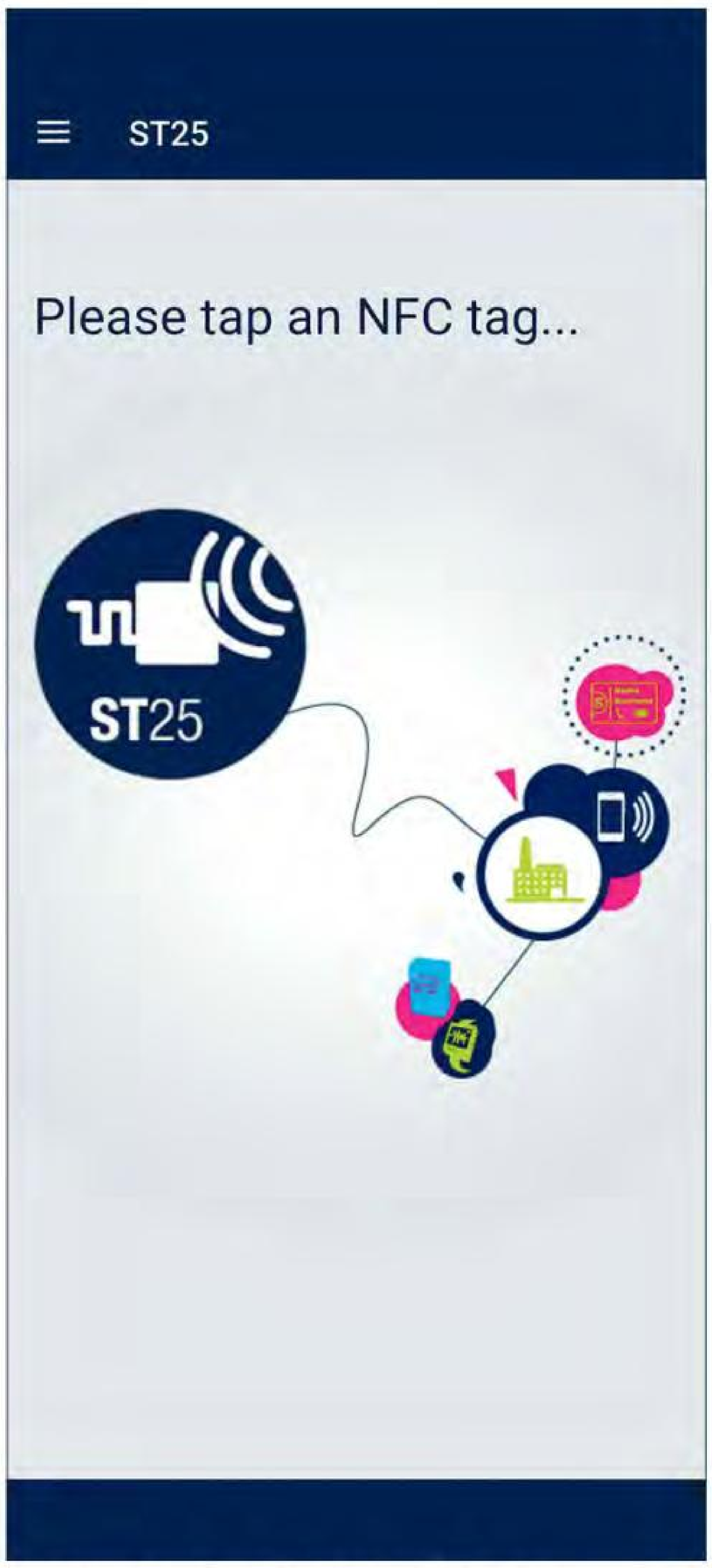
| Protocol | Code | Addr. bits | Cmnd. bits |
|---|---|---|---|
| NEC | N | 8 | 8 |
| Philips RC5 | 5 | 5 | 6 |
| Philips RC6 | 6 | 8 | 8 |
| Sony 12-bit | S | 5 | 7 |
| Sony 15-bit | T | 8 | 7 |
| Sony 20-bit | U | 13 | 7 |

Tap the NDEF tab to see Screen 3, then the blue button at bottom right and select the option to add a plain NDEF text record to the tag. Screen 4 shows the text field; you simply enter the codes and values as shown, pressing the Enter key between each line.

The red button is on the first line, blue on the second and black on the third. When you have finished making changes, save the new text to the tag using the save button at top tight. The ST25 app uses a line feed (LF, ASCII 0xA) as the line separator, so if you use a different app, make sure that this is the same.

The values shown in Screen 4 are equivalent to the default settings provided by the Remote Control. You can also add an extra column with notes or comments about each line. Just be sure to separate it from the other values with a comma and be aware of the limited memory available.

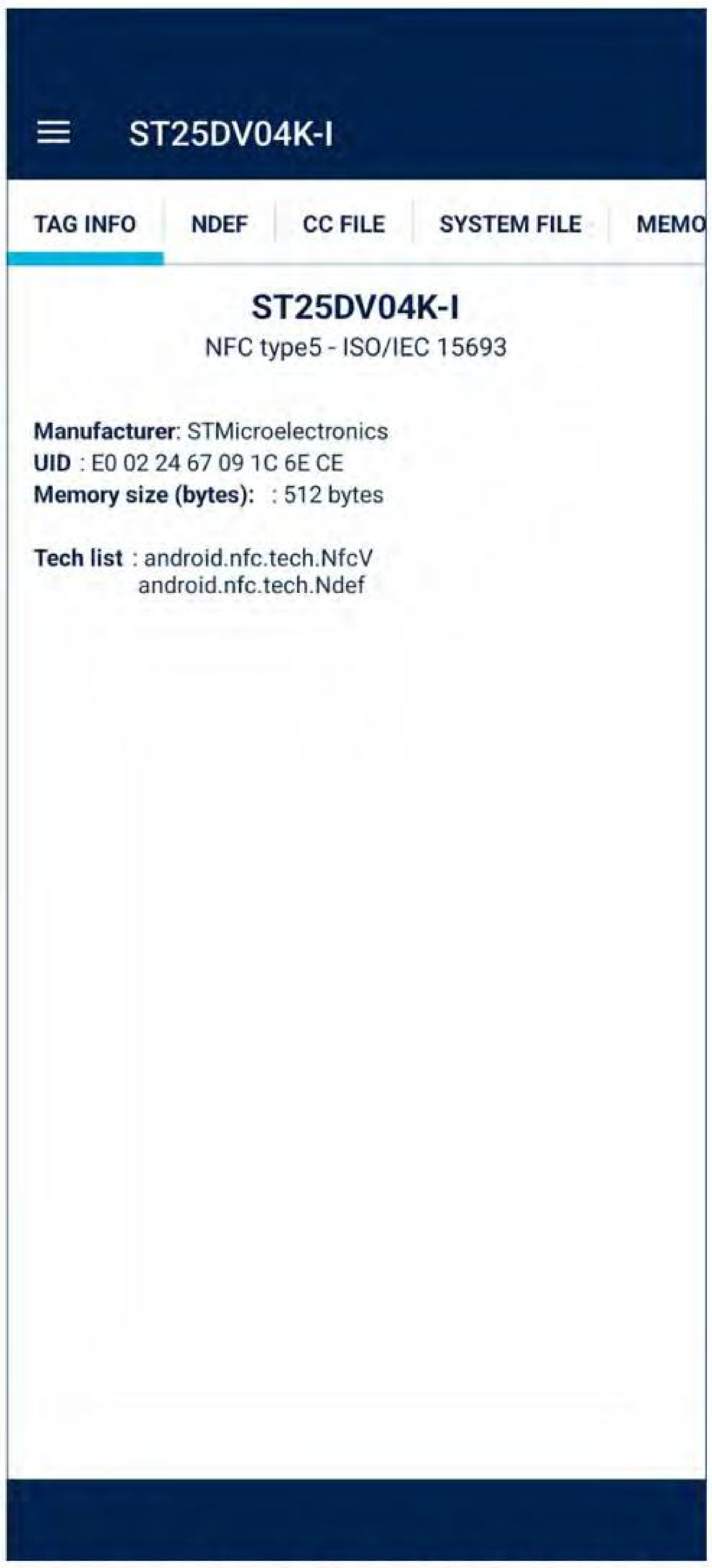
There are many other things that you can do with the app. For example, Screen 3 has a copy button at top right that can be used to clone tags.



ST25

Please tap an NFC tag...

**Screen 1: the ST25 NFC Tap comes from STMicroelectronics, who produce the ST25 range of chips. It's a good idea to open the app before scanning a tag. Otherwise, your device might open a different app when the tag is brought near.**

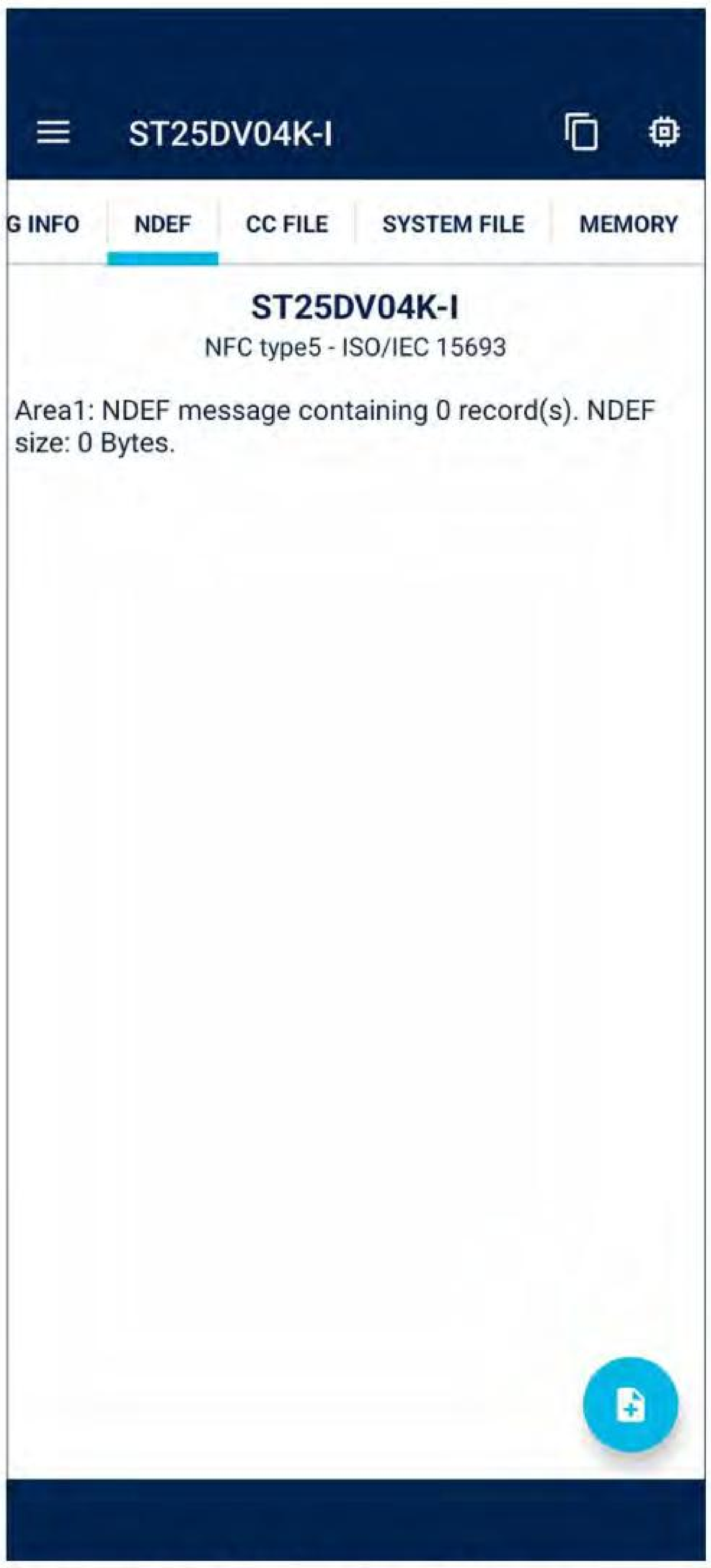


ST25DV04K-I
NFC type5 - ISO/IEC 15693

Manufacturer: STMicroelectronics
UID : E0 02 24 67 09 1C 6E CE
Memory size (bytes): : 512 bytes

Tech list : android.nfc.tech.NfcV
android.nfc.tech.Ndef

**Screen 2: when a tag is first scanned, some basic information is provided, including the serial number. The tabs along the top provide more options.**



ST25DV04K-I
NFC type5 - ISO/IEC 15693

Area1: NDEF message containing 0 record(s). NDEF size: 0 Bytes.

**Screen 3: a blank tag will have no NDEF records yet. The blue button at the bottom right of this page allows records to be added.**

The Memory tab can be used to read, write or erase the tag's EEPROM. If a tag is not working correctly, you can try erasing the EEPROM and rewriting the settings.

The Memory tab also allows the tag contents to be read from or written to a file. The DEFAULT.BIN file in the downloads for this project can be written to the tag to similarly reset it to containing the default IR codes.

You can add other NDEF records to the tag. In our experience, a device will typically act on the first valid record that it recognises. So if you wish to add an alternative record for people to scan (such as a WiFi handover record or URI record), we suggest adding it before the text record for the Remote Control. Non-text records are simply ignored by the microcontroller.

## Remote Control use

While the above process writes a set of codes to IC1, these are not automatically loaded. Instead, the buttons are used to do this under user control. Pressing all three buttons at the same time will trigger the read sequence. While the buttons are held down, the LED will alternate red and green.

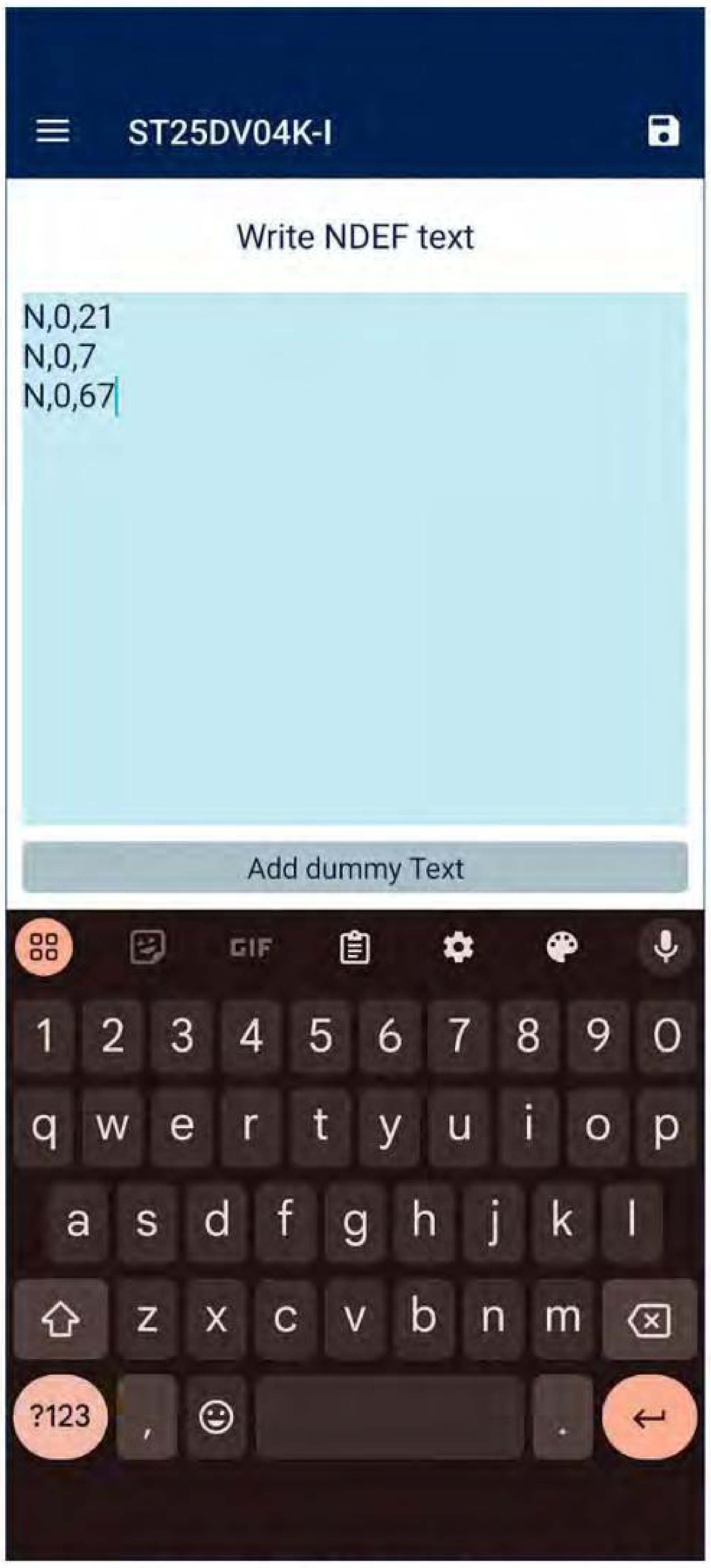
Releasing the buttons starts the reading process. Firstly, IC2 checks if IC1 is present and if it is not, the LED flashes red once for about a second. If IC1 is present but no NDEF text record is found, then nothing is shown on the LED. This can be expected with a blank NFC chip, such as if construction has just been completed.

If a valid NDEF text record is found, the LED will flash once for each of the three button codes, in order from left to right, green if it is valid or red if it is not. If it is valid, the code will be updated; otherwise, the current code is kept. After this, normal operation resumes and you should see a brief flash indicating the battery status.

In general, the code requires a valid protocol code as per Table 1. The address and command values provided must fit within the number of bits prescribed. For example, a value of 256 for either the address or command of an NEC code would be invalid, since these are eight-bit values and 256 requires nine bits to encode.

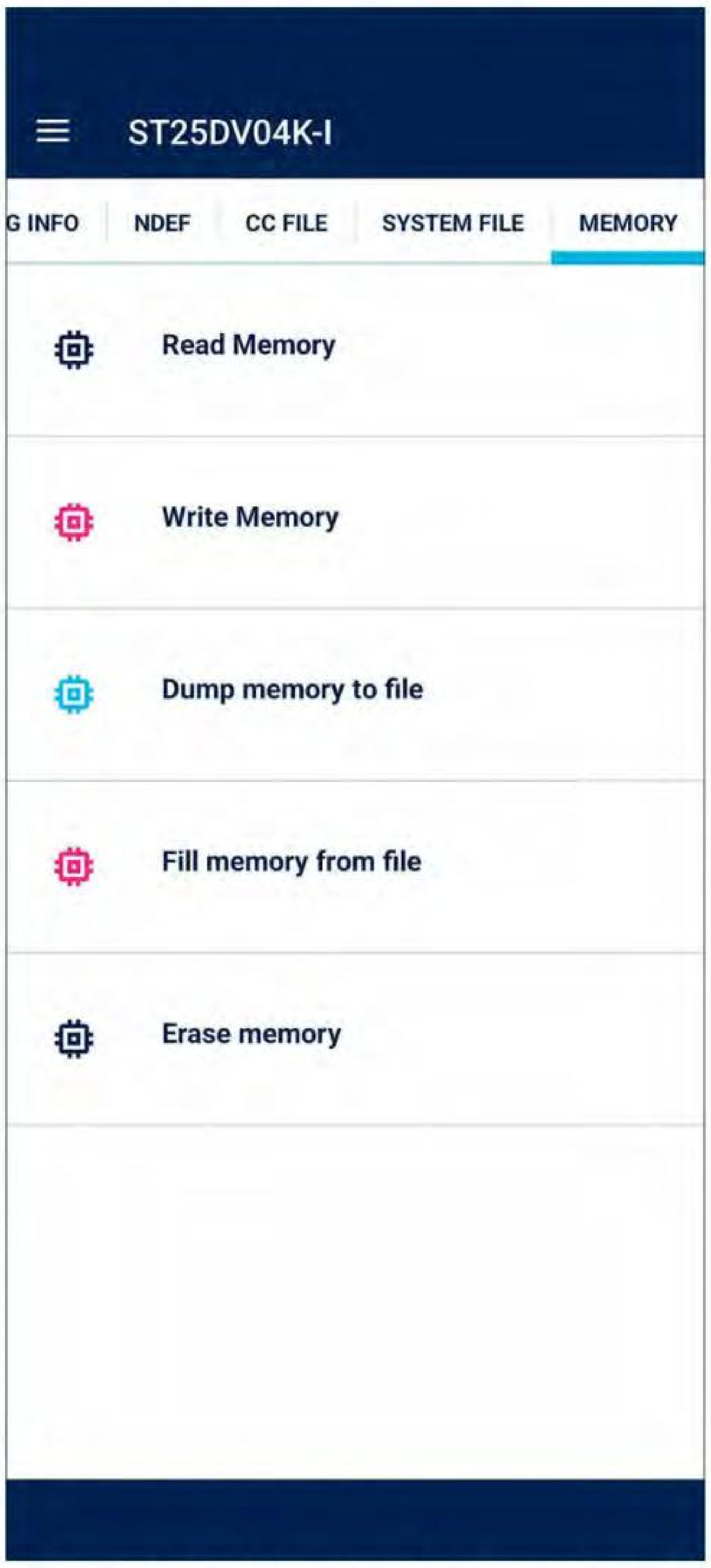
After programming, the Remote Control operates with the new codes. Simply push each button and the corresponding IR code will be sent for as long as the button is held down. If a second button is pressed, while the first is still down, the first code will continue until the first button is released, then the second code will start.

If you start to see the LED flashing red instead of green during operation, then the battery is getting low; down around 2.6V. The circuit itself will function down to near 2.0V, but IR range will suffer due to the lower current provided to LED2.



N,0,21
N,0,7
N,0,67

**Screen 4: the text shown here matches the default settings of the remote control. The text can be stored to the tag with the SAVE button (floppy disk icon) at upper right.**

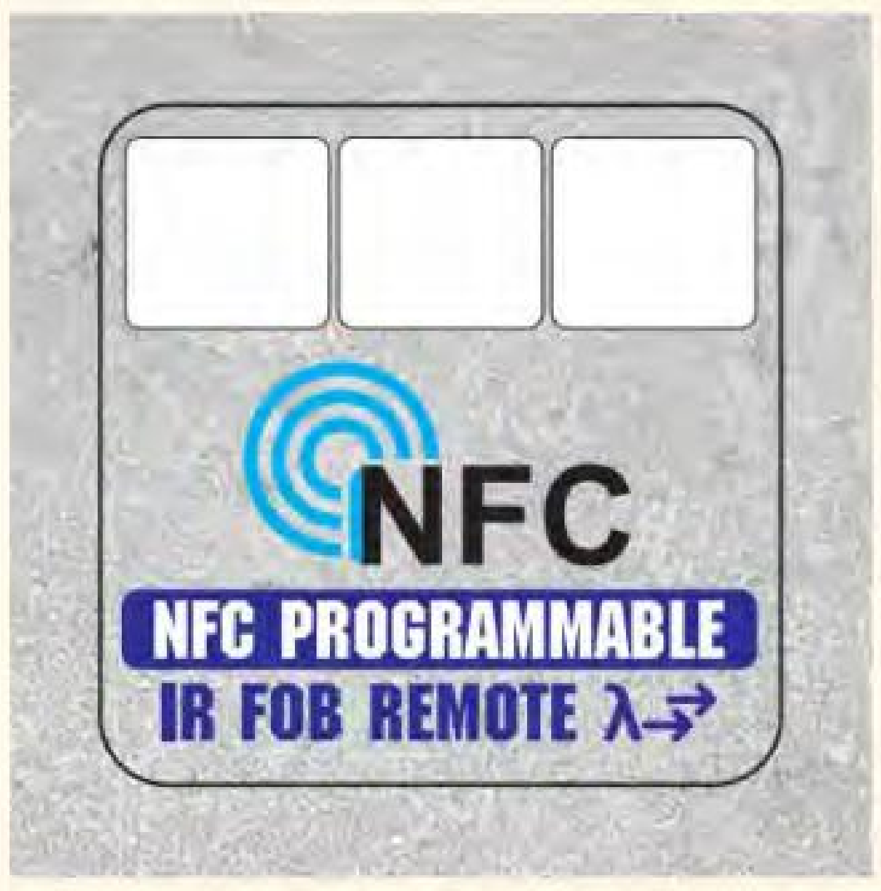


**Screen 5: the Memory tab provides access to low-level read and write functions. You can also store the tag contents to a file.**

### Completion

We've created a label for the keyfob shown below. There's space for functions to be added in permanent marker below each button on the sticker. The kit for this project will include a sticker with this artwork – attach it to the front of the fob case.

As we noted, the files in downloads include a DEFAULT.BIN file (containing the values seen in Screen 4) that can be written directly to the EEPROM, if you wish to experiment with it. The downloads also include the HEX file for programming the microcontroller and the MPLAB X project files.



*There are spaces on this label to add a legend for each button so you know what it does. This will be provided as a sticker in kits purchased from the Silicon Chip Shop, and will also be available to download.*

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# Circuit Surgery

Regular clinic by Ian Bell

## Measuring the frequency response of a circuit using a PC sound card, part 4: analysing circuit frequency responses

**In recent articles, we have** been investigating how to measure frequency responses. This was motivated by the need to measure the response of a practical implementation of an example digital filter (from our series on DSP) without using advanced lab test equipment.

We covered the general principles of frequency response measurement and the use of a "sound card" (audio interface) in a PC or laptop, together with free software called REW (Room EQ Wizard).

Last month, we considered the signal conditioning that might be needed to interface the circuit under test to the sound card's line inputs and outputs. We described part of a signal conditioning circuit built from op amps, the function of which was to get the right signal amplitude and DC levels in and out of the sound card and the microcontroller board implementing the digital filter.

In the circuit description, we mentioned that it is common practice to place a low-value capacitor across the feedback resistor of an op amp amplifier (see Figs.1 & 2). This is primarily to improve stability (reduce the likelihood of unwanted oscillations), one potential cause of which is capacitive loading of the output. We will discuss this further in a future article.

### Low-pass filtering

I also commented that this reduction of gain may help if there are unwanted high frequencies in the input, but more usually an RC low-pass filter and/or ferrite bead would be included in the input circuit to address this if needed.

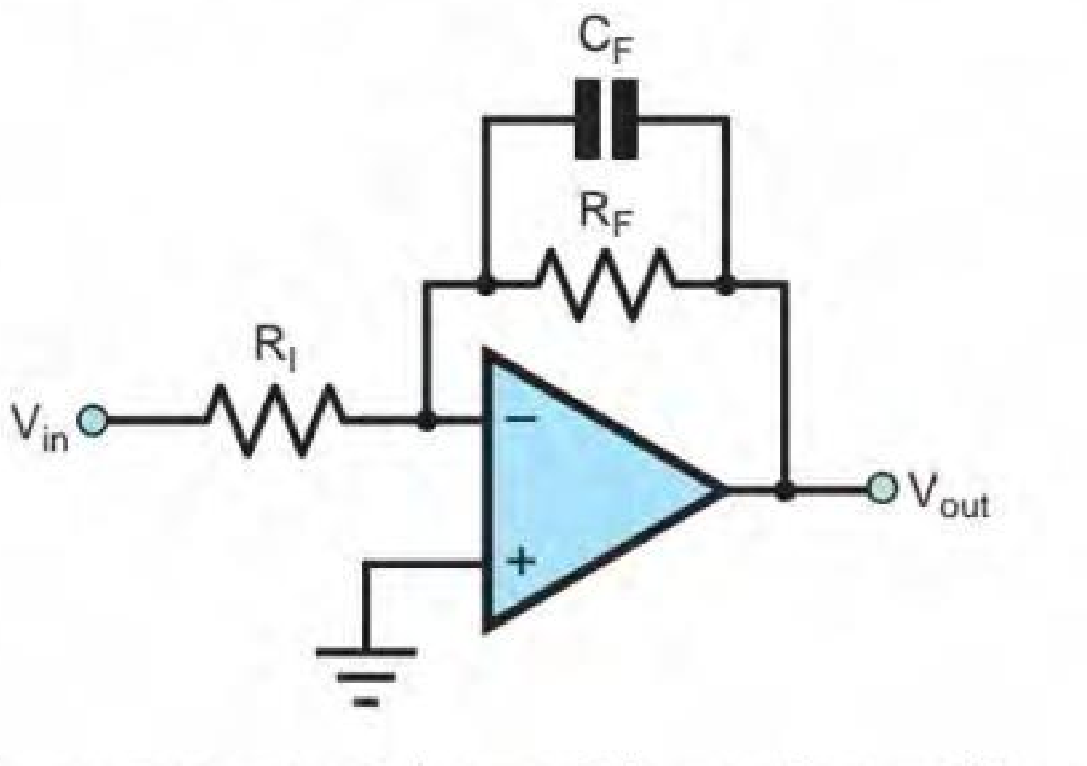

Fig.1: an op-amp-based inverting voltage amplifier with a feedback capacitor.

I was describing a non-inverting amplifier (Fig.2); in this case, the gain does not continuously decrease at higher frequencies, but levels off at unity – so the circuit is not very useful as a low-pass filter in most circumstances.

Despite this, I considered including a formula for the cutoff frequency of the circuit with the capacitor shown in the article. I did not know it from memory, although I knew it was not simply $1 \div (2\pi R_F C_F)$. A quick online search gave me the wrong answer, and various links related to the wrong circuits or circumstances, so I did not include it in the previous articles.

Later, I worked out the formula myself. This highlighted the fact that sometimes when dealing with circuits, it is useful to be able to derive formulae for their characteristics rather than assuming you can look them up. So this article will focus on how to go about such calculations.

I was intrigued and concerned that Google's AI search summary presented the wrong answer. I expect that this is because the circuit is not commonly used as a low-pass filter, so this is relatively little discussed. But this could be misleading to anyone forgetting the small print "AI responses may include mistakes".

### A tale of four AIs

I wondered whether other AI systems would provide better answers. I tried the full AI mode on the Google search site, Copilot on my PC, plus Perplexity and ChatGPT via their websites. I used the same prompt for each, which stated it was a non-inverting amplifier and that I wanted the -3dB frequency, specifically not the pole frequency (poles will be explained later in the article).

I checked my formula and the ones from the AIs against LTspice simulations with idealised op amps. Only ChatGPT gave me a correct response, also providing a detailed derivation. Its final equation was in a similar form to the one I derived manually.

Both Perplexity and Copilot produced a formula with the same general 'look' as the correct versions (a $1 \div 2\pi RC$ term multiplied by a square root term containing the resistor values or gain value), but these were not mathematically equivalent and did not produce the correct numerical values.

Perplexity gave some explanation and source links. Copilot's response was more minimal, so I asked about its sources, and it said its response was based on its internal understanding of analog circuit theory (which was unfortunately not quite good enough). While there were poor results from three out of the four AI systems I tried, they will probably improve in the future.

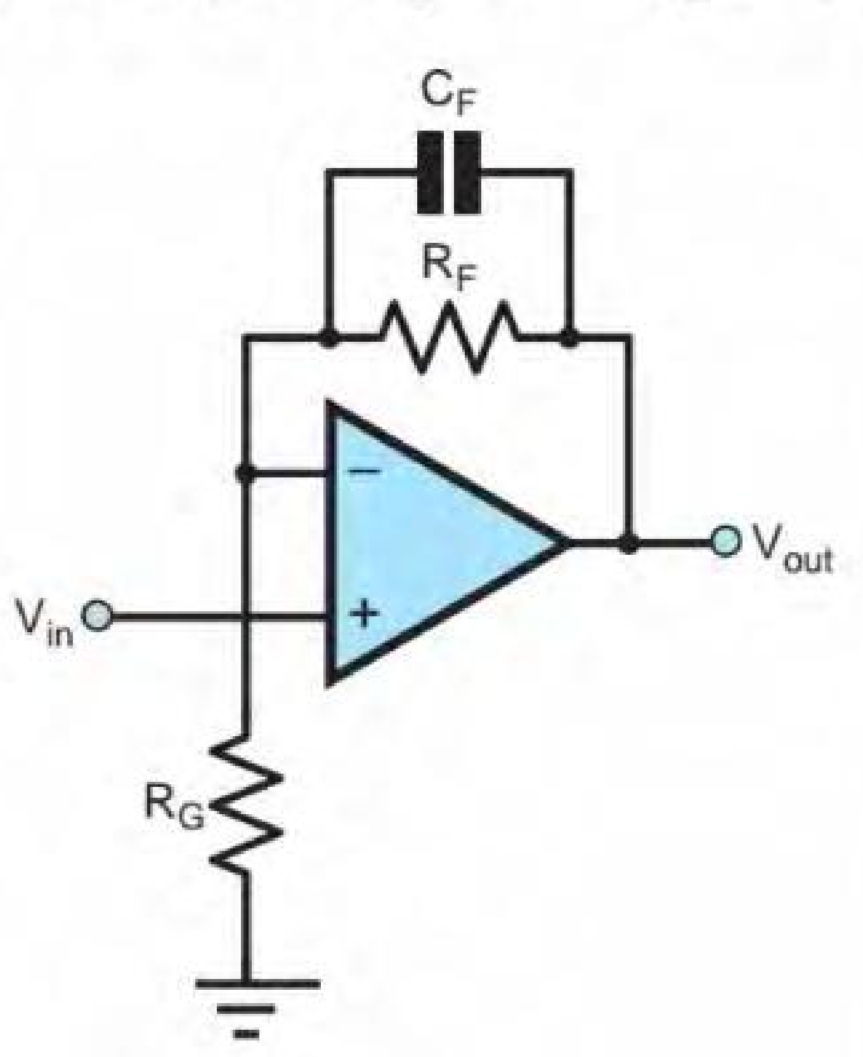

Fig.2: an op-amp-based non-inverting amplifier with a feedback capacitor.

Using AI to obtain circuit equations is potentially very useful but could also undermine the deeper understanding that can come from working through the maths yourself. This month we will go into detail on how to obtain the formula for the -3dB cutoff frequency of the circuit in Fig.2.

This will make use of some techniques that are based on advanced maths, but in many situations only basic algebra and a little knowledge of complex numbers is required. The circuit itself may not be very useful as a filter, but it does provide a good example for discussing how to approach analysing a circuit.

This is also a natural extension of our discussion so far on concepts related to circuit frequency response. We will discuss how to handle the frequency-dependence of capacitors and inductors in circuit calculations.

We will also look at the concept of poles and zeros, which you will often see referred to when discussing issues such as op amp frequency response, and stability of circuits with feedback (eg, in technical documents from semiconductor manufacturers).

## Circuit overview

Having a general or intuitive understanding of a circuit's behaviour before attempting a mathematical analysis is helpful. The equations should fit within the general understanding. For example, if you know that increasing a particular parameter (component value etc) should reduce some other parameter (gain, frequency point etc), the structure of the equation should correspond with this.

The capacitor in the circuits in Figs.1 & 2 causes the circuit to act as a low-pass filter because the capacitor's reactance (effective resistance) reduces as the signal frequency increases.

At low frequencies, the capacitor is effectively an open circuit and has little impact on the gain. As the frequency increases, the capacitor's reactance magnitude becomes comparable with the feedback resistor value, reducing the total parallel impedance. At still higher frequencies, the capacitor dominates the parallel impedance, which tends towards zero as the frequency increases to infinity.

Last month, we discussed the differences between the inverting and non-inverting amplifier configurations. The impact of a feedback capacitor on the amplifier's frequency response is another area where these circuits differ.

For the inverting amplifier, the gain without the capacitor is $-R_F \div R_I$, which is also the DC gain with the capacitor ($A_0$). At higher frequencies, as $C_F$ bypasses $R_F$, reducing the effective overall value of the feedback resistance, the gain of the circuit with the capacitor reduces and tends towards zero (assuming all components are ideal).

For the non-inverting amplifier, the gain without the capacitor (and DC gain with it) is $1 + R_F \div R_G$. At higher frequencies, as $C_F$ bypasses $R_F$, the gain with the capacitor reduces and tends towards unity (the $R_F \div R_G$ term approaches zero, leaving just the 1 term).

The inverting amplifier with a feedback capacitor has a standard first-order low-pass response, whereas the non-inverting amplifier exhibits a shelving filter response. A shelving filter provides different gains at low and high frequencies, but with flat gain-versus-frequency on both sides of the transition region, rather than the gain falling towards zero at the high or low end.

The sinc correction filter discussed in the November 2024 issue has a shelving response. Shelving filters are also common in audio equalisation applications.

The circuits in Figs.1 & 2 are not commonly used as low-pass filters because, with the addition of an extra capacitor and resistor, you can create second-order (faster roll-off) filters using topologies such as Sallen-Key and multiple feedback.

## Complex numbers recap

In the October 2025 issue, we discussed frequency response in general, and briefly discussed complex number representation. Complex numbers are often used in the representation of frequency-dependent circuits, signals and components.

As a reminder, we need to deal with the fact that signals have both amplitude and phase; components and circuits may influence both aspects of a signal. Therefore, a single number is insufficient to represent, for example, the input-output relationship (gain or transfer function) of a circuit, except in simple cases, such as circuits using only resistors.

We can represent amplitude $A$ and phase $\Phi$ directly as $A\angle\Phi$ (see Fig.3), but this does not lead to the most straightforward mathematical analysis of circuits. Therefore, we often use complex numbers where the value of the signal in Fig.3 is $x + jy$ (j is $\sqrt{-1}$, often written as i by mathematicians).

In $x + jy$, $x$ is the 'real' part and $jy$ is the 'imaginary' part (since $\sqrt{-1}$ does not exist as a real number).

It is well known that if we apply a sinewave voltage to a capacitor or inductor, the current is 90° phase shifted with respect to the voltage. Thus, the current-voltage relationship of capacitors and inductors and circuits built from them requires the use of either an amplitude-phase or complex number approach.

## Reactance and impedance

For resistors, the current ($I$) to voltage ($V$) relationship is Ohm's law, $V_R = RI_R$, where all the values are real numbers. Capacitors and inductors have a similar relationship, with reactance ($X$) rather than resistance. For example, $V_C = X_C I_C$ for a capacitor. More generally, for circuits with a mixture of components, we use impedance ($Z$), so $V = ZI$. Reactance and impedance involve complex numbers.

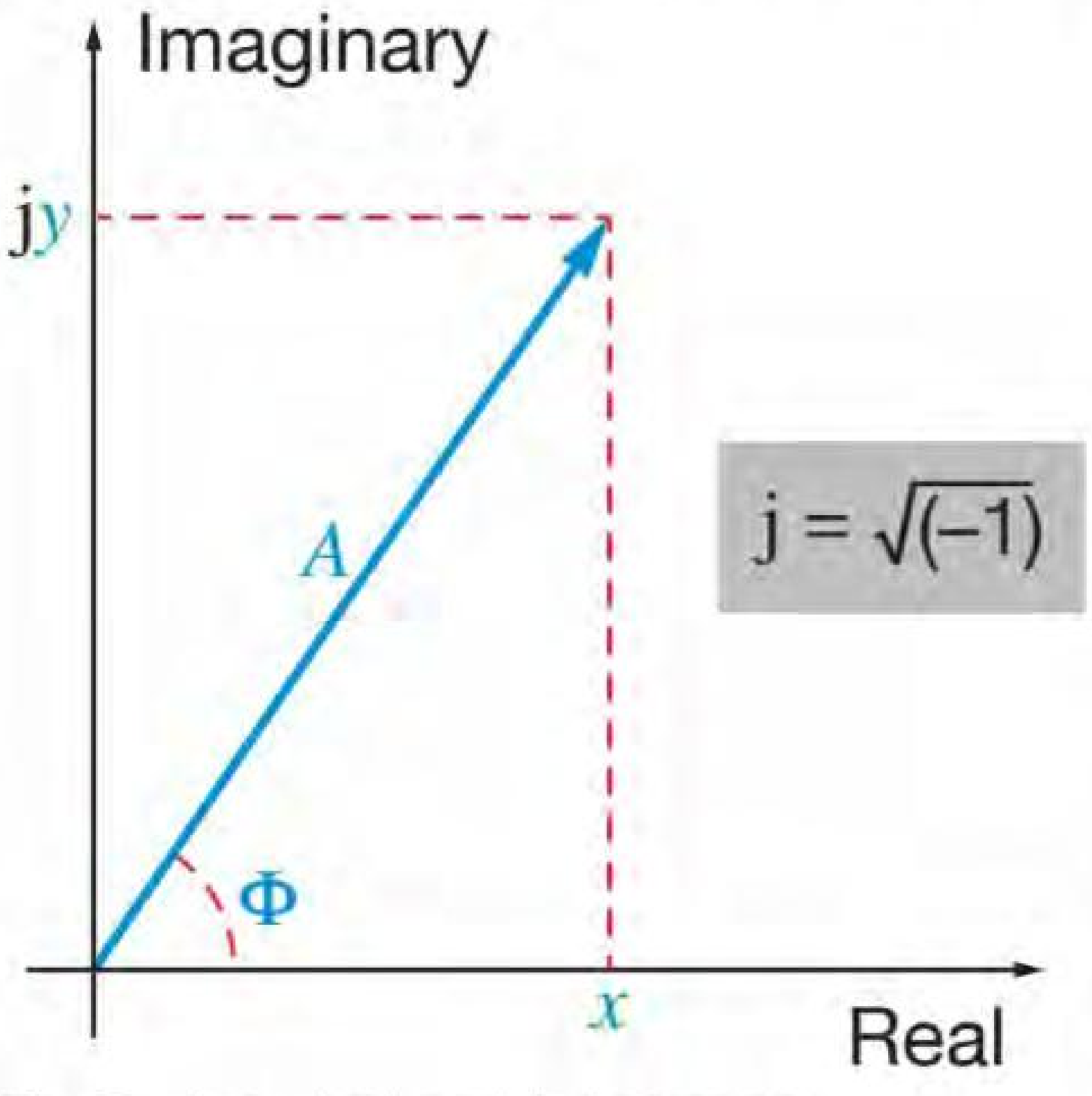

Fig.3: the relationship between amplitude, phase and complex numbers.

The mathematics underlying circuit analysis with complex numbers will not be familiar to all our readers. However, in some common cases, it is possible to use complex impedances for analysing circuits, along with well-known rules such as Ohm's law, series and parallel component combinations and the potential divider formula, without a deep knowledge of advanced maths.

The following is an informal 'hand-waving' explanation of the complex reactance of capacitors and inductors.

As frequency increases, the reactance of a capacitor decreases – it is inversely proportional to frequency (proportional to $1/f$). Also, at a given frequency, increasing the capacitance decreases the reactance; again, it is inversely proportional. We can calculate it as $1 \div 2\pi fC$ with frequency in hertz ($f$) or $1 \div \omega C$ with frequency in radians per second ($\omega$), but we also need to consider the phase shift.

The 90° phase shift means that the voltage-to-current relationship for the capacitor must be represented as a purely imaginary number (the real and imaginary axes are at 90° in Fig.3).

The fact that voltage lags current in a capacitor means that in $V_C = X_C I_C$, the value of $X_C$ is purely imaginary and negative, giving $X_C = -j/\omega C$. Multiplying by $-j/-j$ gives $X_C = -1/-j\omega C = 1/j\omega C$, which is the form you often see it written in.

Similar formula manipulations for an inductor (where voltage leads current and impedance is proportional to inductance and frequency) gives $Z_L = j\omega L$.

## Transforming problems

Fundamentally, the instantaneous current ($i$) in a capacitor ($C$) is proportional to the rate of change of voltage ($v$) across it. We can write this as:

$$i = C\frac{dv}{dt}$$

Here, $dv/dt$ is the rate of change of $v$ with respect to time ($t$) – we are in the realm of calculus. Similarly, for an inductor, we have $v = L(di/dt)$. The resistor is simpler – its time domain equation is Ohm's law, $v = Ri$.

For circuits containing resistors, inductors and capacitors, we can use these three component characteristic equations, along with circuit theory rules, to develop equations relating input signals to output signals. The results are time-domain differential equations. Unfortunately, such equations are difficult to directly manipulate and solve.

To overcome this difficulty, we use mathematical transforms. Transforms convert problems between different domains (the way in which signals are represented), for example, the time domain or the frequency domain. This can greatly simplify things; for example, requiring solutions to algebraic rather than differential equations.

The most commonly used transforms in electronics are the Fourier and Laplace transforms for continuous signals, and the Z transform for sampled signals. Others, such as the wavelet transform, are also used in specific circumstances where they provide advantages.

Developing the theory, proofs and properties of transforms from scratch takes a significant amount of advanced maths, but this has already been done. So, in straightforward cases, we do not have to directly apply the transform maths. We can use the known results, like the component reactances ($1/j\omega C$ and $j\omega L$) discussed above, together with 'resistor-based' circuit theory.

Of course, we need some knowledge of manipulating complex numbers. Often, this just involves finding the magnitude of an impedance, or a gain-magnitude for a frequency response calculation (remember it is the gain magnitude that is plotted in a Bode frequency response graph).

The magnitude of a complex number, $x + jy$, is $\sqrt{x^2 + y^2}$. This follows from applying Pythagoras' theorem to find $A$ in Fig.3. When finding the magnitude of a division of two complex numbers, say $M \div N$, we find the individual magnitudes of $M$ and $N$ first, then perform the division.

For purely capacitive and inductive reactance, we only have an imaginary part (eg, $jy$), so the magnitude is $\sqrt{y^2} = y$. For example, this is $1 \div \omega C$ or $1 \div 2\pi fC$ for a capacitor, which is well-known as the formula for the 'resistance' of a capacitor.

It might look like we just removed j to get $1 \div 2\pi fC$, but if we have a combination of resistance and reactance, this does not work. We have to use the full root-sum-squared calculation to get the impedance magnitude – as we did last month when we looked at the cutoff frequency of the AC-coupled inverting amplifier.

## Enter the *s*-domain

The Fourier transform represents signals in the frequency domain ($f$ or $\omega$), based on the fact that signals can be represented by a sum of sinusoids. The Laplace transform uses the *s*-domain, which is a form of complex-number frequency.

The Laplace transform is a generalisation of the Fourier transform, which can handle signals that the Fourier transform can't (when you delve into the transform maths), so it is more widely applicable. In the *s*-domain, the capacitor and inductor reactances are $1/sC$ and $sL$, respectively.

The *s*-domain is able to handle a wide range of signals and situations, but if we limit ourselves to the steady-state, single-frequency behaviour of linear circuits, we can use the substitution $s = j\omega = j2\pi f$. We discussed steady-state conditions previously in our introduction to frequency response concepts (October 2025).

Typically (for steady state), we can perform the algebra of our circuit analysis using *s*-domain component reactances, make the substitution for $s$ in the result, then find the magnitude to get the real numbers we need. We could do all the calculations using $j\omega$ or $j2\pi f$ instead of $s$; you will find examples of both approaches in technical documentation.

A common approach to circuit analysis is to use the *s*-domain component reactances to obtain the transfer function $H(s)$ in the form of a division of two polynomials in $s$, with the denominator $A(s)$ and numerator $B(s)$; that is:

$$H(s) = \frac{B(s)}{A(s)}$$

A polynomial of a variable (eg, $s$) is an expression involving only the variable and its non-negative integer exponents (powers), which are multiplied by coefficients (dimensionless positive or negative numbers) and summed, eg, $s^2 + 2s + 5$.

## Poles and zeros

The values of $s$ that cause $B(s)$ and $A(s)$ to be zero indicate important frequencies in the circuit's transfer function, determining the shape of the frequency response. When the numerator $B(s)$ is zero, the *s*-domain transfer function (gain) value is zero, and this is given the obvious name of a zero.

When denominator $A(s)$ of the *s*-domain transfer function value is zero, the transfer function approaches infinity – this is called a pole (a pole sticking up; a high point in the values).

Putting $s = 0$ in the transfer function gives the DC gain. For example, this transfer function has one pole and one zero:

$$H(s) = \frac{1 + (1/628)s}{1 + (1/4712)s}$$

The zero is found using $1 + (s \div 628) = 0$ (numerator equals zero), so $s \div 628 = -1$, hence $s = -628$. If we substitute $s = j\omega = j2\pi f$, we get $f = -628 \div (j2\pi)$. To find the frequency, we calculate:

$$f = \frac{\sqrt{(-628)^2}}{\sqrt{(2\pi)^2}} = \frac{628}{2\pi} = 100\,\text{Hz}$$

There is a zero at 100Hz. The pole is found using $1 + s \div 4712 = 0$ which, following a similar calculation, gives 750Hz. The DC gain ($s = 0$) is 1.

Remember that $s$ is a complex (two-dimensional) value that is related to frequency but is not frequency as we experience it. An *s*-domain transfer function going to infinity does not mean the real gain magnitude of the circuit is infinite. Having poles that are at negative values of $s$ does not mean negative real frequencies.

The abstract nature of the *s*-domain and advanced maths behind it can be off-putting, but it is a useful tool for circuit analysis calculations. Furthermore, poles and zeros provide important insights into frequency response and circuit stability.

At frequencies above the pole frequency, gain decreases at a rate of 20dB per decade more than at frequencies below the pole. This is a change in the rate-of-change of gain with frequency, not the absolute rate of change, so if the gain was decreasing by 20dB per decade below a pole, it would decrease by 40dB per decade above it.

A pole also shifts the phase by -90°. Fig.4 shows the frequency response characteristics of a circuit with a single pole.

Similarly, for a zero, the gain will increase at a rate of 20dB more per decade above the zero frequency compared to below it. A zero adds +90° of phase shift. Fig.5 shows the frequency response characteristics of a circuit with a single zero.

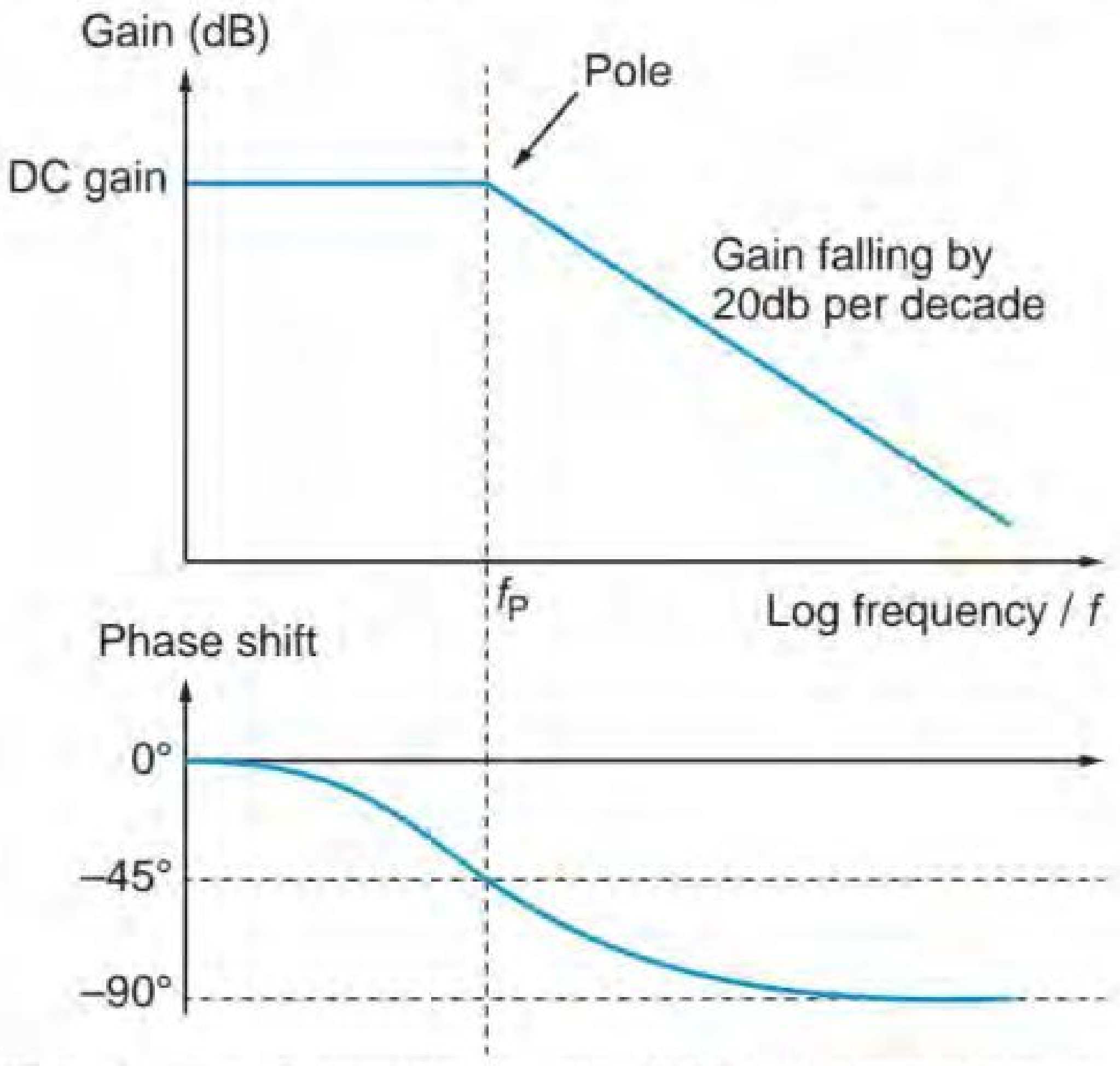

Fig.4: the frequency and phase response of a circuit with a single pole.

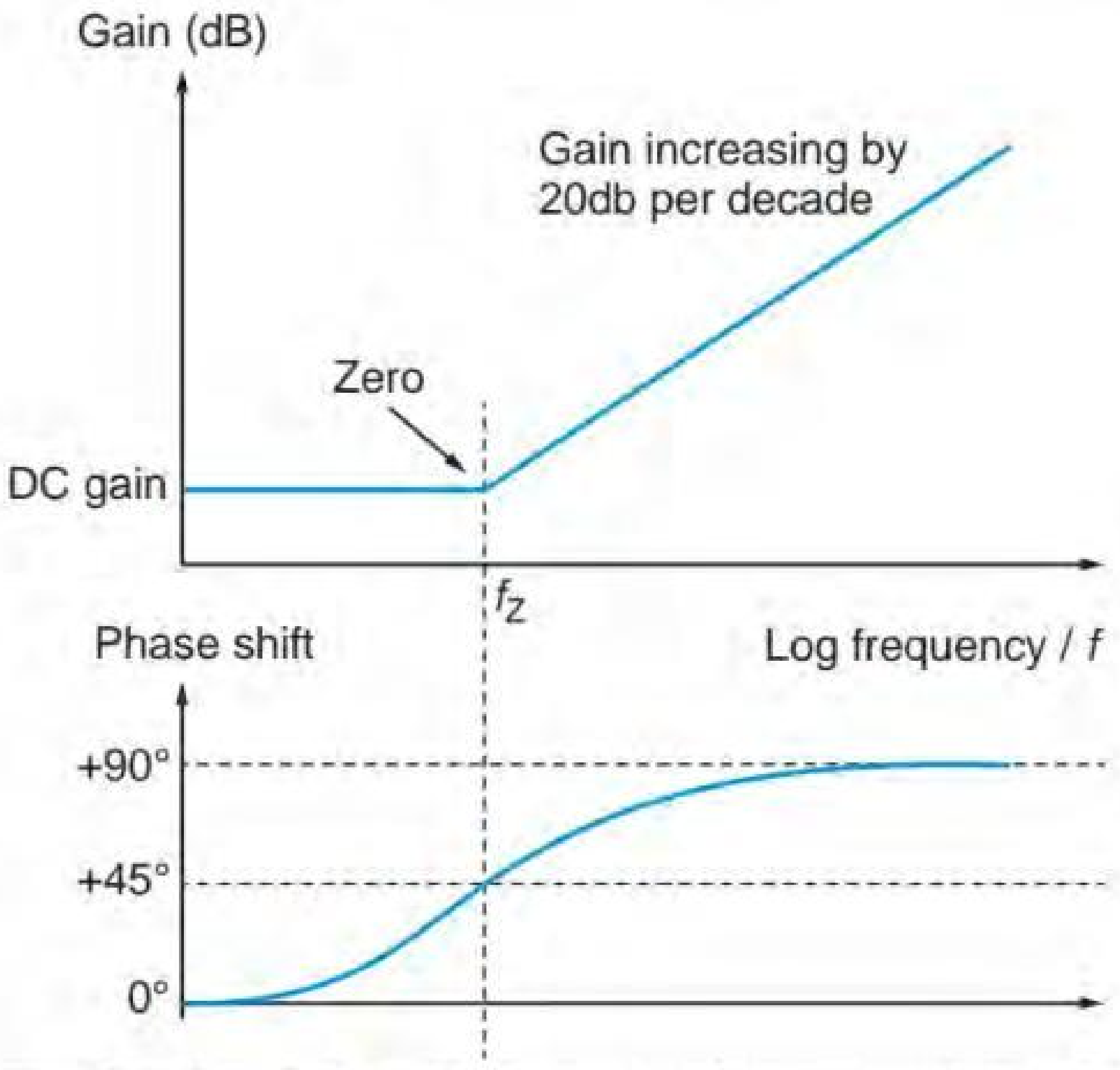

Fig.5: the frequency and phase response of a circuit with a single zero.

## LTspice Laplace

LTspice can directly implement Laplace functions using controlled and behavioural sources. Fig.6 shows a voltage-controlled voltage source (VCVS, E SPICE element) implementing a similar example to the one above:

$$H(s)=\frac{1+\left(1/6.28\times 10^{7}\right)s}{1+\left(1/628\right)s}$$

Here, the pole is at 100Hz, the zero is at 10MHz and the DC gain is 1. Very different pole and zero frequencies were chosen for this example so that the effect of each on the frequency response is well separated and clear to see.

The VCVS is configured using **`Laplace=<func(s)>`** as the source "value" (its gain). The keyword **`Laplace`** is followed by a function (**`func`**) in terms of ***s***. The above equation in LTspice format is **`Laplace=(1+s/6.28e07)/(1+s/628)`**.

Fig.7 shows the results from the simulation in Fig.6. We can see that the breakpoints in the gain magnitude response (solid green trace) occur at 100Hz and 10MHz, corresponding to the pole and zero frequencies. The plot also shows the phase response (dotted green trace and right-hand axis).

As the frequency increases past the pole, the slope of the gain response decreases by 20dB per decade from 0dB per decade (horizontal) to -20dB per decade. The pole causes the phase shift to change by -90°, with -45° of the phase change occurring at the pole frequency. The specific change here from 0° to -90° takes place over a range of frequencies above and below the pole frequency.

As frequency increases past the zero, the slope of the gain response increases by 20dB per decade from -20dB per decade back to 0dB per decade (horizontal again). The zero causes the phase shift to change by +90°, with +45° of phase change occurring at the zero frequency. The specific change from -90° back to 0° again takes place over a range of frequencies.

The DC (or low-frequency) gain is ×1 (0dB), which is the gain at 1Hz seen in Fig.7.

Fig.8 shows a simulation using the circuit in Fig.6 with the zero frequency changed to 500Hz. The closeness of the pole and zero means that the full effect of the pole (-90° phase and -20dB per decade slope) does not occur before the zero reverses the situation.

## RC circuits

The discussion so far has been somewhat abstract, so we will look at a simple circuit example – an RC low-pass filter, shown in Fig.9 along with a Laplace version of its transfer function implemented using a behavioural voltage source (B1). The cutoff frequency is well known for

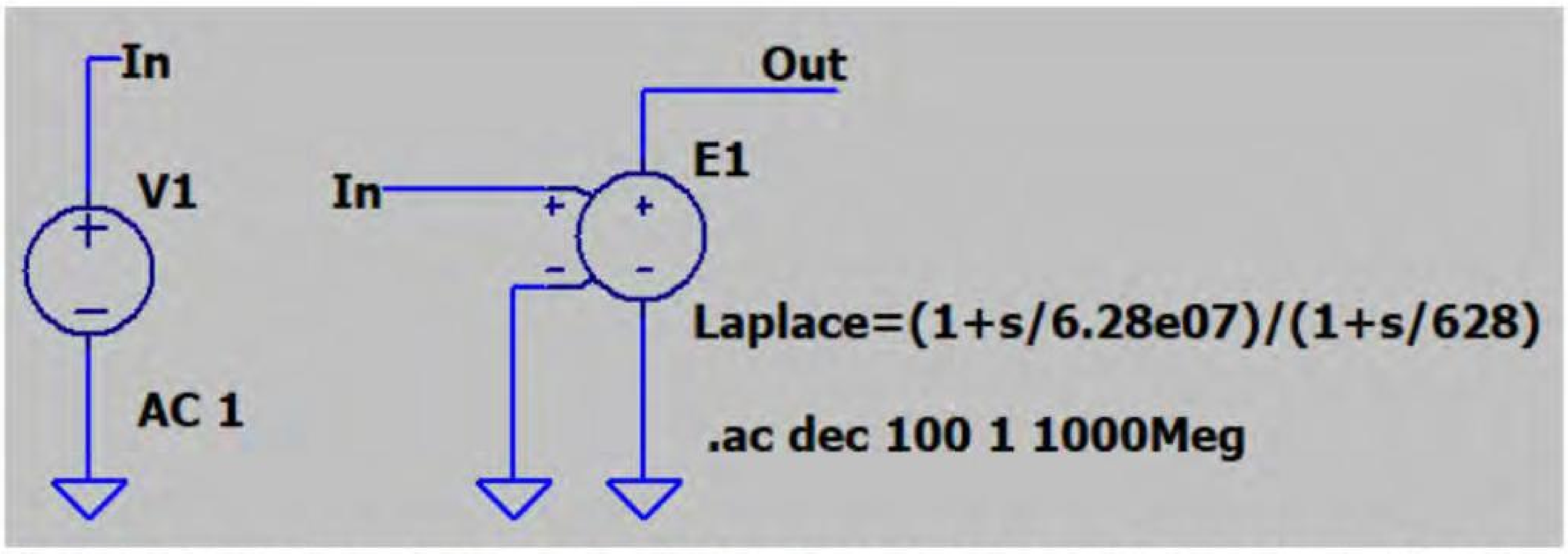

Fig.6: implementing a Laplace transfer function using a controlled source in LTspice.

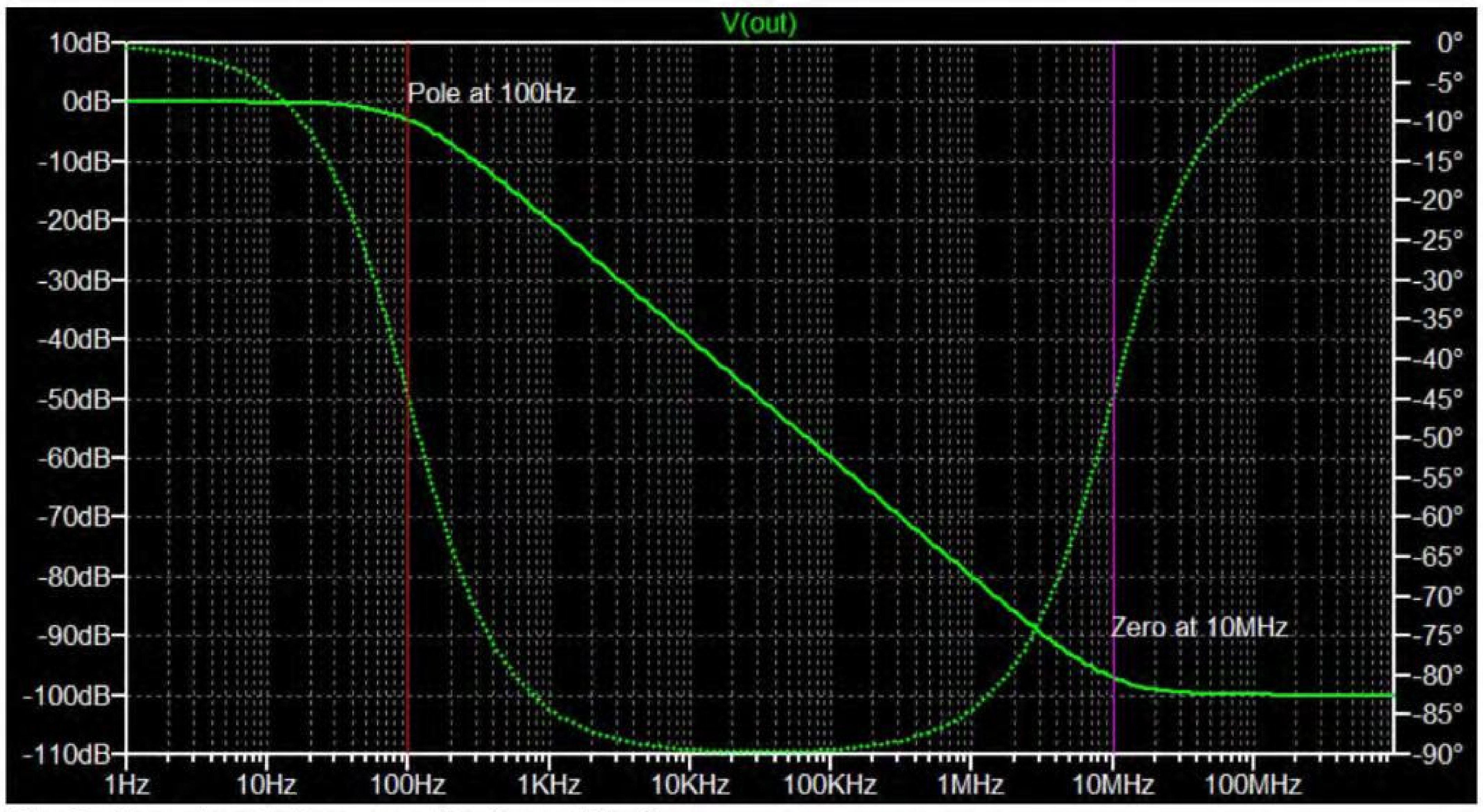

Fig.7: the simulation results from Fig.6.

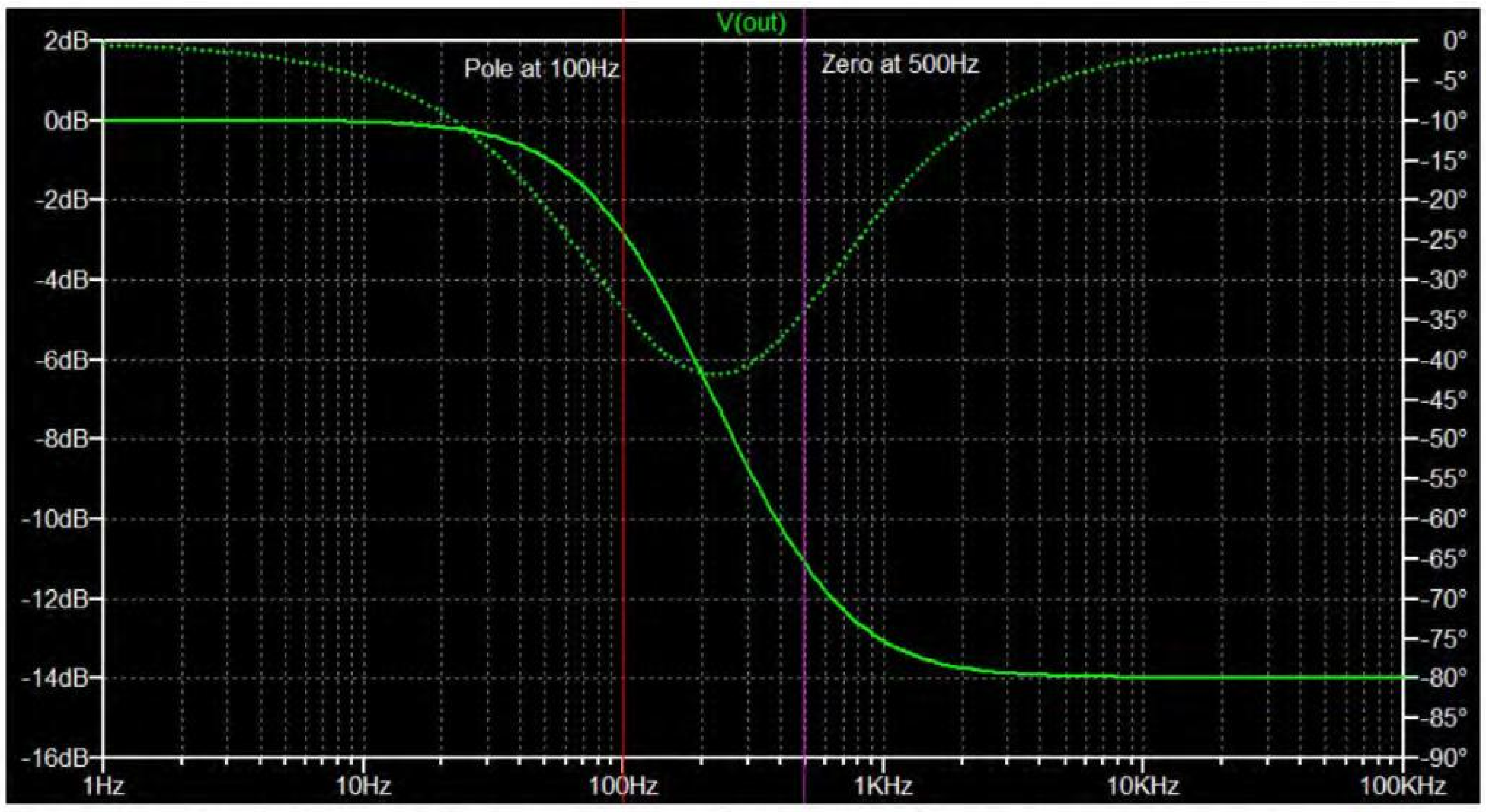

Fig.8: the simulation results from Fig.6 with closer pole and zero frequencies.

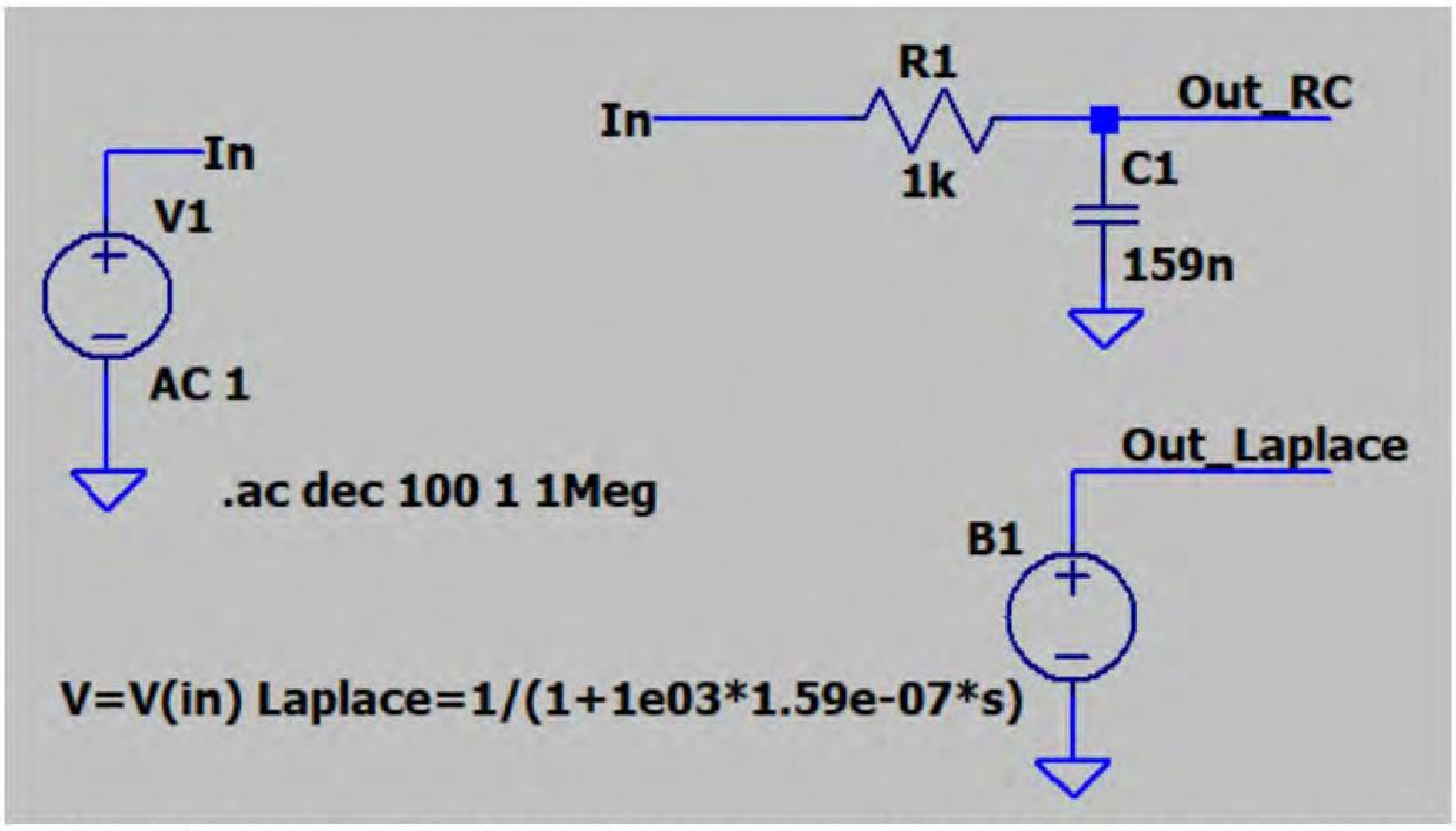

Fig.9: an LTspice simulation for the Laplace representation of an RC low-pass filter.

this circuit; $f = 1 \div 2\pi RC$, which with the values used here is $1 \div (2 \times \pi \times 1000 \times 159{\times}10^{-9}) = 1.0$kHz.

The circuit is a potential divider formed by a resistor ($R$) and capacitor ($C$). In the $s$-domain, the resistor's resistance is simply $R$ and the capacitor's reactance is $X_c = 1 \div sC$, as previously discussed. Using the potential divider formula, the output is:

$$v_{out}(s) = \frac{X_c}{R+X_c} v_{in}(s) = \frac{1/sC}{R+1/sC} v_{in}(s)$$

Note that we write $v_{in}(s)$ because we are working in the $s$-domain and this applies to the voltages as well as the component values. Multiplying the top and bottom by $sC$ and moving $v_{in}$ to get the expression for gain, $H(s)$, in the polynomial form, we obtain:

$$H(s) = \frac{v_{out}(s)}{v_{in}(s)} = \frac{1}{1+sRC}$$

This response has no zeros (there is no $s$ in the numerator). It has a pole at $1 + sRC = 0$, meaning $s = -1/RC$. If we substitute $s = j\omega = j2\pi f$ and find the magnitude:

$$f = \frac{\sqrt{(-1)^2}}{\sqrt{(2\pi RC)^2}} = \frac{1}{2\pi RC}$$

This follows the same pattern as the numerical example above (s = -628). We are back to the well-known -3dB cut-off frequency of the circuit: $f = 1 \div 2\pi RC$. The pole frequency and -3dB frequency are the same here, but in general, you cannot assume this is the case.

The negative value of s is important in wider use of $s$-domain analysis, but here it disappears when we take the magnitude to find the pole frequency.

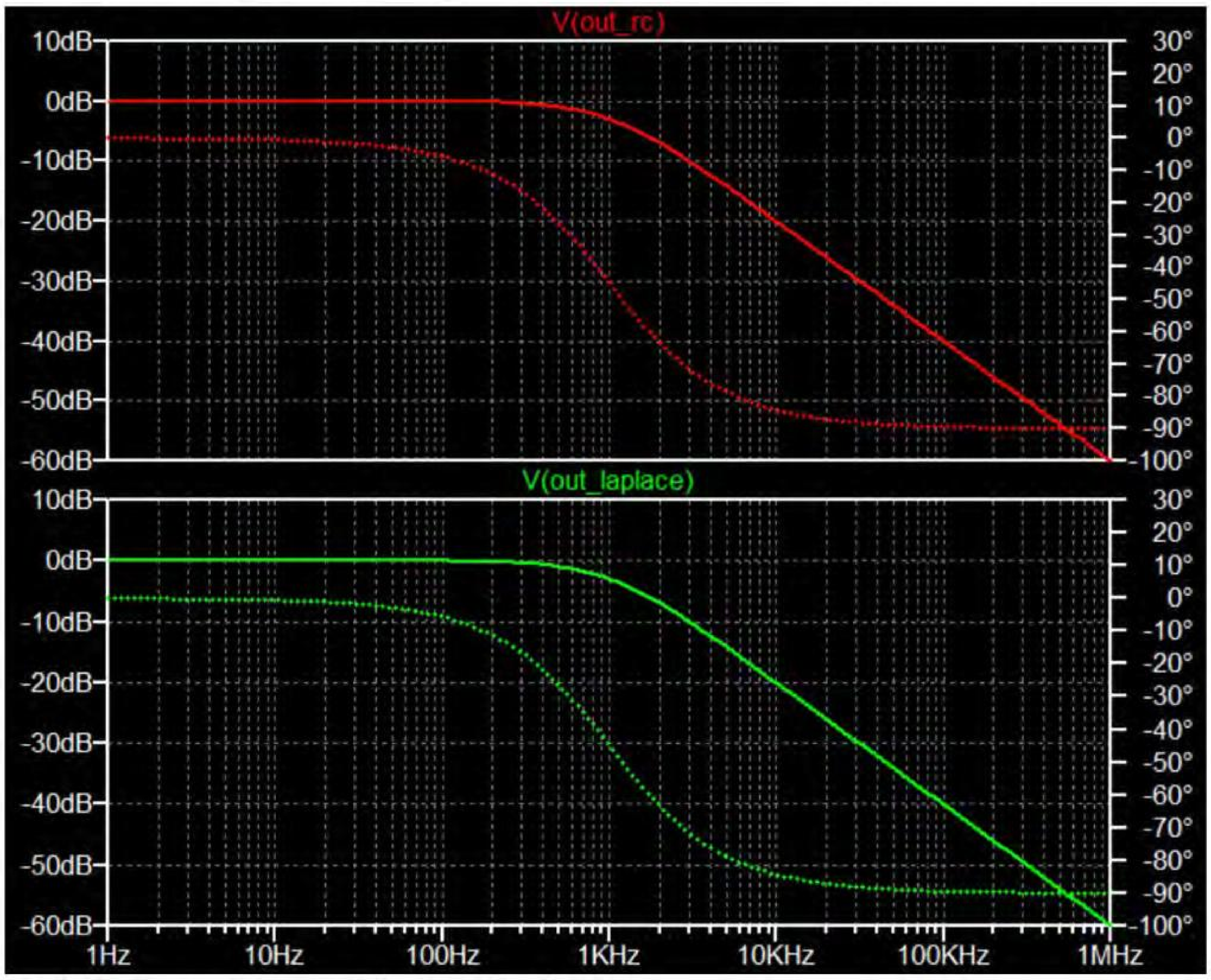

Fig.10: the simulation results from Fig.9.

The syntax for implementing the Laplace function using an LTspice behavioural source is a little different from the controlled sources such as the VCVS used above. If the Laplace function definition is included in the source definition, it is applied to what would usually be the source output, as defined by the source expression, **`V=<expression>`**.

In this case, the expression is just **`V=V(in)`**, so the source is applying the voltage on node **`in`** to the Laplace function, which is what we need here. The Laplace statement is written after the source expression, separated by a space, and is the $H(s)$ function for the RC circuit from above with the specific R and C values in it: **`Laplace=1/(1+1e03*1.59e-07*s)`**.

The results from the simulation shown in Fig.10 confirm that the RC circuit and Laplace function have the same frequency response.

This example illustrates that with the $s$-domain reactances, basic circuit theory, and a little knowledge of complex numbers we can obtain and use the Laplace domain transfer function without having to do any difficult calculations, such as applying the transform maths directly to a circuit equation.

It's worth emphasising that a pole or zero will always indicate a breakpoint in the frequency response, but this will not necessarily be a 3dB point. The effect of poles and zeros can overlap, which leads to more complex relationships between the poles and zeros and gain at these frequencies.

In fact, if there is a pole and a zero at exactly the same frequency, the two breakpoints are present, but their effects cancel one another and no change in gain or phase will occur.

Poles and zeros are fundamental in terms of defining the shape of the frequency response. A 3dB point is an arbitrary (but commonly used) definition of a cut-off frequency. It is not used universally.

For example, a Chebyshev filter's cutoff frequency is commonly defined as the frequency after the last passband peak where the response has dropped by the ripple level, which is not necessarily at -3dB.

## Non-inverting amplifier response

For the non-inverting amplifier, the gain without the capacitor and (also DC gain with the capacitor) is $A_0 = 1 + (R_F \div R_G) = (R_G + Z_F) \div R_G$. With the capacitor in place, we need to replace $R_F$ with the parallel impedance ($Z_F$) of $R_F$ and $C_F$, which is the parallel combination of $R_F$ and the capacitor's reactance, $X_{CF}$. So, the gain is $1 + (Z_F \div R_G)$.

The parallel resistance ($R_P$) of multiple resistors ($R_1$, $R_2$, $R_3$ and so on) is given by the well-known formula:

$$\frac{1}{R_P} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \ldots$$

For two resistors, this becomes $R_P = (R_1R_1) \div (R_1 + R_2)$. As well as resistors, we can use the Laplace impedances of the capacitor and inductors in these formula, so we get:

$$Z_F = \frac{R_F X_{CF}}{R_F + X_{CF}} = \frac{R_F(1/sC_F)}{R_F + (1/sC_F)}$$

Multiplying both the numerator and denominator by $sC_F$, we get:

$$Z_F = \frac{R_F}{1+sR_FC_F}$$

Putting this expression for $Z_F$ into the gain equation, written as $(R_G + Z_F) \div R_G$, we get:

$$H(s) = \frac{R_G + R_F/(1+sR_FC_F)}{R_G}$$

Multiplying both the numerator and denominator by $1 + sR_FC_F$, we get:

$$H(s) = \frac{R_G + R_F + sR_FR_GC_F}{R_G + sR_FR_GC_F}$$

This form of the equation allows us to find the poles and zeros. Note that we get the DC gain (the gain of the non-inverting amplifier without $C_F$) as expected with $s = 0$.

There is a pole at $R_G + sR_FR_GC = 0$, at which $s = -R_FC$ (the $R_G$ terms cancel). Using $s = j\omega = j2\pi f$ gives the pole frequency as $f_p = 1 \div 2\pi R_FC$, which is the same as the inverting amplifier.

There is a zero at $R_F + R_G + sR_FR_GC = 0$, which gives us:

$$f_z = \frac{R_G + R_F}{2\pi R_F R_G C} = \frac{1}{2\pi R_F C} A_0$$

$A_0$ is the DC gain of the amplifier. In comparison, the inverting amplifier does not have a zero in its transfer function – its gain decreases continuously with frequency past the pole frequency.

## Finally, the -3dB frequency

To find the -3dB point, we have to find the frequency at which the magnitude of the gain is 3dB below the passband gain (the DC gain in this case). As discussed last month, this is when the gain is reduced by a factor of $\sqrt{2}$, so we need to find the frequency ($f_{\text{-3dB}}$) at which $|H(s)| = A_0 \div \sqrt{2}$.

The algebra to solve this (and similar problems) could be quite 'messy' – there are many terms to keep track of, and it could take quite a few steps. The individual operations are not difficult or advanced (add, subtract, multiply and square root to manipulate the equations), but it is easy to make a mistake, and not necessarily obvious what the most effective approach is.

It is possible to manipulate the equation in different ways to get to the result. Of course, when you see the finished working presented nicely in an article or book, you don't know if a took a couple of attempts to get a slick approach. Don't worry if you try similar calculations and things don't go totally smoothly.

One way to approach this particular example is to see that the DC gain formula can be found within the expression for $H(s)$. If we divide the numerator and denominator by $R_G$, we get:

$$H(s)=\frac{(R_G+R_F)/R_G+sR_FC_F}{1+sR_FC_F}$$

$$H(s)=\frac{A_0+sR_FC_F}{1+sR_FC_F}$$

We can also make a temporary substitution of $P = sR_FR_GC$, where $P$ indicates this is the term associated with the pole and the parallel resistor and capacitor. We do this just to reduce the number of terms we have to write out. It also helps us see that the equation comprises terms related to the DC gain and the pole. Having done this, we can write the -3dB gain equation $|H(s)| = A_0 \div \sqrt{2}$ as:

$$\frac{\sqrt{A_0^2+P^2}}{\sqrt{1+P^2}}=\frac{A_0}{\sqrt{2}}$$

We need to solve this for $P$ (make $P$ the subject of the equation). This takes a few steps: square both sides, cross multiply, then gather and cancel terms, arranging so that all terms in $P^2$ are on one side. We get:

$$P^2=\frac{A_0^2}{A_0^2-2}$$

We can now reinsert the original terms in $P$, make the $s = j2\pi f$ substitution and make $f$ the subject:

$$f_{-3dB}=\frac{1}{2\pi R_F C}\cdot\frac{A_0}{\sqrt{A_0^2-2}}=\frac{f_p A_0}{\sqrt{A_0^2-2}}$$

We could also substitute the resistors back into the $A_0$ terms, but this form of the equation is more useful for understanding the circuit as it more directly shows the relationship between the -3dB point, the pole frequency and the DC circuit gain.

## What the formulae tell you

When you obtain an equation for a circuit, you can look at the relationship between terms to gain insights into the characteristics of the circuit. Earlier, we noted that the inverting and non-inverting circuits differed in their pole frequencies.

The inverting amplifier analysis was straightforward, and non-inverting case far less so. The + 1 term in the non-inverting amplifier gain ($1 + R_F \div R_G$) is responsible for this.

A commonly used approach to using formulae to understand circuits is to look at what happens when key terms in the expression corresponding to characteristics of the circuit (or signal) become very large, or very small, or take on specific values that have a significant impact on the way the equation works.

We are looking for situations where a formula can be simplified – where values calculated by the formula are dominated by some terms and others become almost irrelevant. We are also interested in situations where the equation can't be solved or produces (or tends towards) specific values like one, zero or infinity.

We will look at some situations where these things happen in the formula for the -3dB frequency of the non-inverting amplifier, together with numerical examples to help make sense of it . Later we will show some related simulation results.

At high DC gains, the – 2 term in $A_0^2 - 2$ has little influence on the value of the denominator; for example, for $A_0 = 50$, the denominator is $\sqrt{50^2-2} = 49.98$. If we ignore the 2, the denominator becomes $\sqrt{A^2} = A = 50$. The difference is small enough (compared to normal tolerances in component values etc) to justify simplifying the equation under these conditions.

If the denominator is simplified to $A_0$, it cancels with the $A_0$ in the numerator and the formula for the cutoff frequency reduces to $f_{\text{-3dB}} = f_p = 1 \div 2\pi R_F C$, which is the same as for the non-inverting amplifier.

At lower values of $A_0$, we cannot make this simplification; for example, for $A_0 = 3$ we have 2.65, which gives $f_{\text{-3dB}} = 1.13f_p$ and hence a 13% error in -3dB frequency if we assume the pole and cutoff frequency are the same.

The term $\sqrt{A^2-2}$ tells us that $A_0^2 = 2$ is significant. At this point, the denominator becomes zero, so the -3dB frequency is at infinity. When $A_0^2$ is less than 2, we have the square root of a negative value.

This formula is based on the gain magnitude, which is a real number; we are not in the realm of complex numbers, so the square root of a negative value indicates that a solution does not exist. This means that when $A_0^2 < 2$, there is no -3dB frequency.

At exactly $A_0^2 = 2$, the gain is $A_0 = \sqrt{2} = 1.414$ (3dB). Remember that the minimum gain of the non-inverting amplifier is 1 (0dB), so at $A_0 = \sqrt{2}$, the maximum gain drop (going from DC to high frequency) is from $\sqrt{2}$ to 1, which is a factor of exactly $1/\sqrt{2}$ (0.7071), that is, exactly a 3dB drop. But that will only happen at an infinitely high frequency.

For DC gains lower than $A_0 = \sqrt{2}$, the drop in gain from $A_0$ to 1 as frequency increases will be by a factor of less than $1/\sqrt{2}$ – there is insufficient gain 'headroom' to accommodate a 3dB drop, so there is no -3dB frequency. For example, at a DC gain of $A_0 = 1.2$ (1.58 dB), the gain can only drop by 1.58dB.

At low gains, the circuit still has a pole and a zero at the frequencies given by expressions derived above. For example, for $A_0 = 1.2$, with $R_F = 200\Omega$, $R_G = 1\text{k}\Omega$ and $C_F = 796\text{nF}$, we have a pole at 1kHz and a zero at 1.2kHz. As the gain reduces, the pole and zero get closer together.

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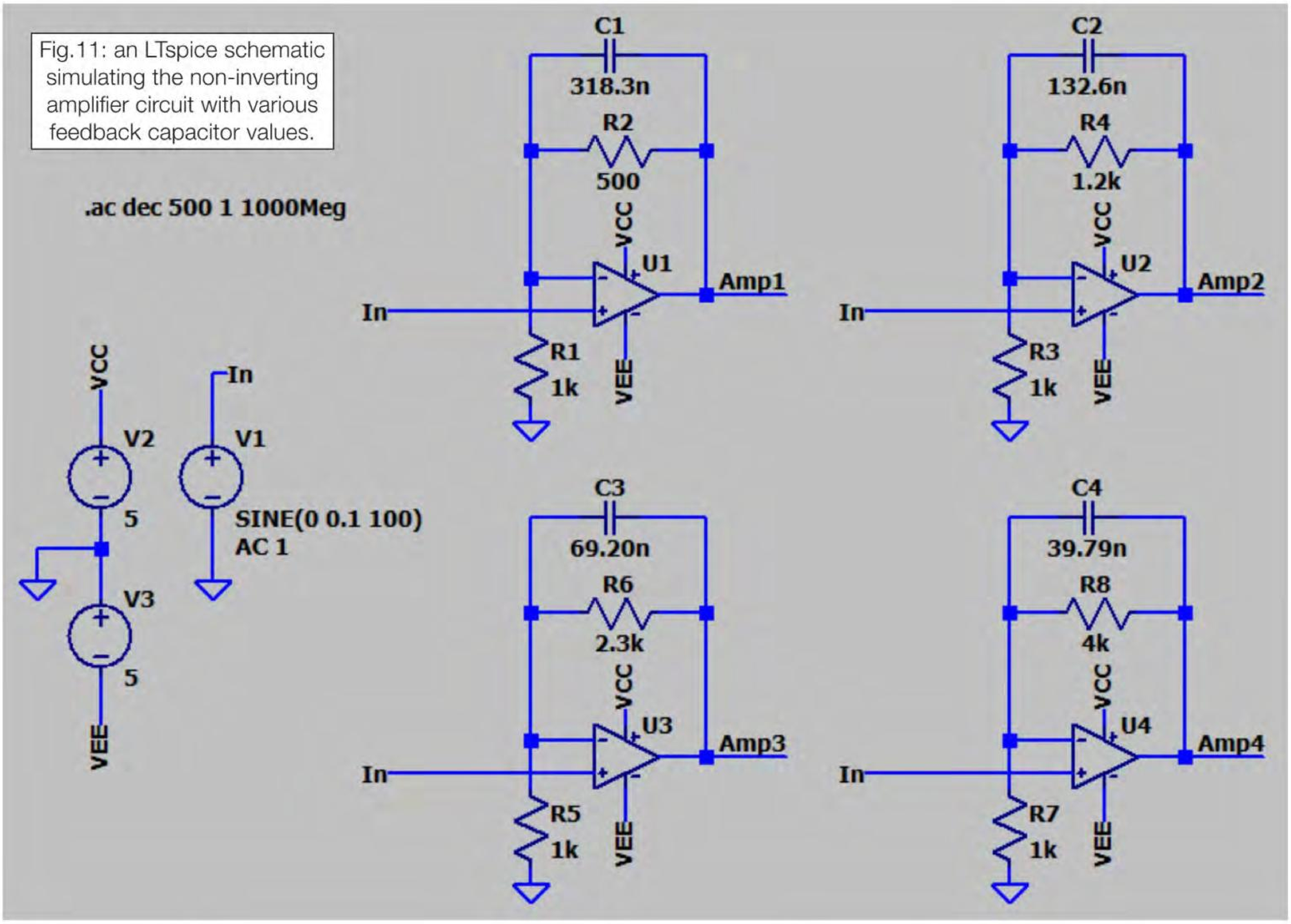

Fig.11: an LTspice schematic simulating the non-inverting amplifier circuit with various feedback capacitor values.

## LTspice examples

Fig.11 shows an LTspice schematic for simulating four copies of the inverting amplifier with different gains but the same pole frequency of $f_p$ = 1kHz. The circuits were configured by setting all the $R_G$ resistors (R1, R3, R5 and R7) to 1kΩ and choosing a set of DC gains ($A_0$ values of ×1.5, ×2.2, ×3.3 and ×5.0) that are approximately evenly spaced in decibel terms, at +3.5dB, +6.9dB, +10.4dB and +14.0dB, respectively. The op amps are the idealised UniversalOpAmp1 model from LTspice library, with the open loop gain and gain-bandwidth product increased from the defaults to more ideal values.

The values of the $R_F$ resistors (R2, R4, R6 and R8) were calculated using the standard non-inverting amplifier gain formula. The capacitor values were calculated by rearranging the pole frequency formula for the circuit (as discussed above) to make the $C$ the subject ($C = 1 \div 2\pi f_p R_F$).

The results are shown in Fig.12, where we see that as the DC gain is reduced, the -3dB point moves further from the pole frequency. The DC gains are approximately evenly spaced, but the shift in -3dB frequency gets larger as the DC gain decreases. This is consistent with the discussion on the formula, which showed the -3dB frequency moving towards infinity as the DC gain gets close to +3dB.

The -3dB frequency values shown in Fig.12 were measured from the simulation. They are very close to the frequencies calculated using the formula derived above. The largest difference is 2.979kHz from the simulation versus the 3kHz calculated value (a 0.7% difference).

This is due to numerical inaccuracies in the simulation, errors in measuring from the graph and possibly slight non-idealities of the op amps. **PE**

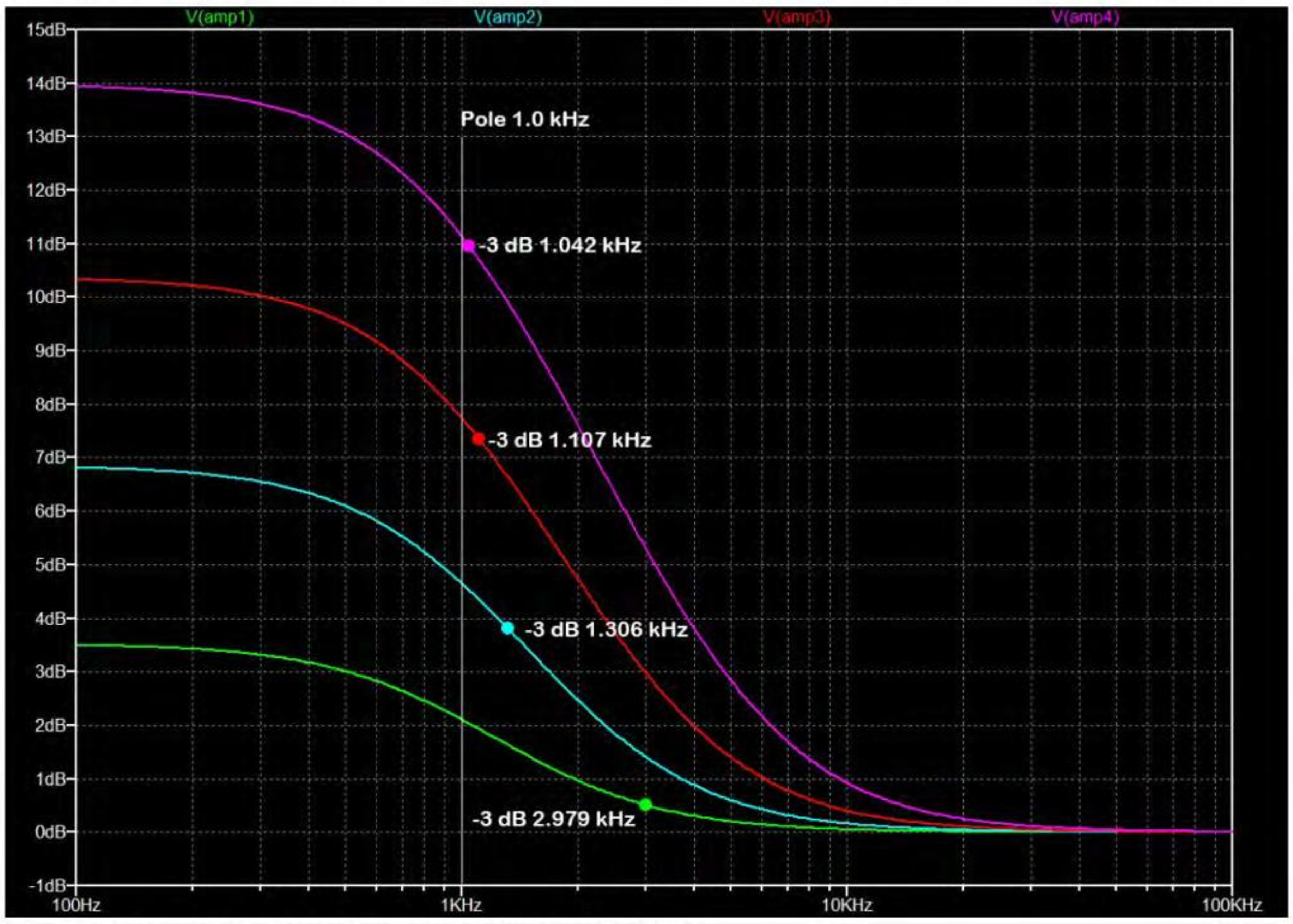

Fig.12: the results from the simulation in Fig.11.

### Simulation files

Most *Circuit Surgery* columns feature the use of the free circuit simulation software LTSpice. It is used to support descriptions and analysis in *Circuit Surgery*.

The examples and files for this issue are available for download: https://pemag.au/link/ac9b

# AOShop

The home for specialist audio, analogue and historic components – provided by Jake Rothman, *PE's Audio Out* columnist. The AOShop is your best bet for classic analogue 'NOS' (new old stock) components, including all parts for *Audio Out* projects and designs.

## Ge semiconductors

*Small-signal PNP transistors*

| | |
|---|---|
| NKT214F, OC57, OC59, 2N1377, 2N525 | £1.00 |

*Low-noise PNP transistors*

| | |
|---|---|
| GET106 | £2.50 |

*Small power PNP transistors*

| | |
|---|---|
| AC153, AC153K, AC188 | £1.50 |

*Small power NPN transistors*

| | |
|---|---|
| AC176, AC141K | £2.00 |

*High-voltage PNP transistors*

| | |
|---|---|
| OC77, CV7001 | £1.50 |

*TO3 PNP power transistors*

| | |
|---|---|
| OC22, CV7054 (OC23), OC25, OC35, OC36, AD143, AD149, AD161, AD162 | £2.00 |
| AD140 | £3.50 |
| AD149, AD161/2 matched pair | £5.00 |

*RF PNP transistors*

| | |
|---|---|
| OC41, OC42, 2SA12, 2SA53,AF124, AF178, GET872A | £1.50 |

*NPN transistors*

| | |
|---|---|
| OC139, OC140, ASY73 | £2.00 |

*Diodes*

| | |
|---|---|
| CV7049 (OA10), CG92 (OA91) | £0.50 |

## Si semiconductors

*Diodes*

| | |
|---|---|
| ZC5800 RF Schottky | £0.20 |

*Low-noise silicon transistors*

| | |
|---|---|
| BFW16A, 2SC3071, 2SC3068, 2SA1016K, 2SC2362K, 2SA970BL | £1.50 |
| 2SC2204, 2SD655, BC550C | £0.50 |
| ZTX651 | £0.30 |
| ZTX751 | £0.50 |

*RF transistors (suitable for Theremin)*

| | |
|---|---|
| BF199 | £0.50 |

*Audio power MOSFETs*

| | |
|---|---|
| Exicon 10N20, 10P20 | £6.50 |
| Hitachi 2SJ99, 2SK343 | £3.50 |
| Hitachi 2SJ56, 2SK176 | £8.50 |

*JFETs*

| | |
|---|---|
| BFW11, BFW10, TIS73L, J177, J113, U1994, U1898, 2SJ176, J201 | £1.00 |
| J175, J176, J112, J111, 2N3820, 2N5467, BF244, 2N5460, J230 | £0.60 |

*Small power output/driver transistors*

| | |
|---|---|
| 2SB649A, D669A, 2SA1208, 2SC2910, MJE253G, MJE243G, 2SA1725, 2SC4511 | £1.20 |
| BD139,BD140, BD135, BD136, BD435, BD436, MPSA63, BCV46 | £0.50 |

*MOSFETs*

| | |
|---|---|
| ZVP2106A | £0.30 |
| Dual-gate 3SK45, BFS28 | £1.50 |

*Metal-cased transistors*

| | |
|---|---|
| BCY71 | £0.30 |

*Amplifiers*

| | |
|---|---|
| LM318 high-speed op amp | £0.35 |
| µA709 metal-cased op amp | £2.00 |
| 5534H metal-cased op amp | £3.00 |
| LM384, TDA2030A, TDA2050V | £1.50 |
| TAA435 (Mullard power amp driver for AD161/AD162) | £1.20 |

## Synthesiser ICs

| | |
|---|---|
| That/dbx 2180 VCA/VCF | £6.00 |
| CA3080 VCA/VCF (vocoder) | £2.80 |
| CA3280 VCA/VCF | £4.00 |
| LM13600/ LM13700 VCA/VCF | £2.00 |
| CA3086 transistor array | £1.00 |
| PT2399 echo/delay | £1.50 |

## Dual transistors

| | |
|---|---|
| 2N2639, 2N2223, 2N2910 (NPN) | £4.00 |
| E401 (JFET Moog) | £4.00 |
| 2SK2145-Y dual JFET | £0.80 |
| 2N5564 JFET | £8.00 |
| HN3C51F, HN3A51F | £1.00 |
| DMMT3904/6, HN1A01F, HN1C01F | £0.50 |

## Loudspeakers

*PE Mini-Monitor*

| | |
|---|---|
| Volt PE165 6.5-inch woofer (each) | £85 |
| Morel MDT29 tweeter (each) | £25 |
| Kit *pair* of PE165/MDT29 plus Volt crossover parts and PCBs | £299 |
| Monacor DT-28N tweeter (each) | £35 |
| Vifa 19mm BC20SC15-04 tweeter (each) | £15 |

*Volt crossover inductors*

| | |
|---|---|
| 1.2mH, 1.5mH, 2mH, 2.7mH, 0.5mH (tapped at 0.3mH) (each) | £5.85 |

*Fully assembled and tested high-quality speaker prototypes – ask for details*

| | |
|---|---|
| LS3/5As and other similar speaker systems (pair) | £200-£350 |

*Fully tested reclaimed speakers*

| | |
|---|---|
| Vifa BC14 5-inch woofer (each) | £10 |
| Vifa TC26 1-inch tweeter (each) | £10 |

*Low-price speakers*

| | |
|---|---|
| Philips 4-inch 4070 | £2.00 |
| EMI 10x6-inch, 30Ω Alnico | £7 |
| 64mm 64Ω neodymium | £1.20 |

## Capacitors

***Note '10/63' denotes '10µF 63V'.***

*Polyester*

| | |
|---|---|
| 3.3/100, 4.7/250, 4.7/63 | £1.00 |
| 5.6/63, 8.2/63, 10/63 | £2.00 |
| Mullard 'Mustard' C296 0.22/400 | £2.00 |

*Polycarbonate*

| | |
|---|---|
| Axial 2.2/63 1%, 4.7/160, 6.8/63 | £1.00 |

*Radial*

| | |
|---|---|
| 6.8/160V, 10/63 | £2.00 |
| Reclaimed 22/63 | £2.00 |

*Polystyrene*

| | |
|---|---|
| Philips 1% 4.7nF/160, 6.2nF/500, 12nF/63, 22nF/63, 110nF/63, 24nF, 2nF | £1.00 |
| RIFA 1% 100nF/100, Suflex 90.9nF 0.5% | £2.00 |
| Suflex 2.5% 10nF/63 (rad. or ax.) | £0.50 |

*Radiation resistant*

| | |
|---|---|
| Siemens cellulose acetate MKL 2.2/25 | £0.80 |

*Electrolytic – Mullard blue 017 series*

| | |
|---|---|
| 10/25, 22/25, 100/10 | £0.50 |
| 150/40, 470/40, 1000/40 | £1.00 |

*Tantalum – axial metal cased*

| | |
|---|---|
| 22/50, 47/35, 68/25, 100/20, 120/10, 150/16, 220/10, 330/6 | £2.00 |
| 22/35, 33/35, 47/20, 68/15, 100/10, 150/6 | £1.25 |
| 4.7/50, 6.8/35, 10/25, 10/35, 22/15 | £1.00 |

*Axial moulded-case tantalum*

| | |
|---|---|
| Kemet axial 6.8/10 | £0.30 |
| Kemet radial 33/10 | £0.40 |
| STC radial 100/20 | £1.50 |

*Tantalum bead*

| | |
|---|---|
| 22/50 | £1.00 |
| 470/3 | £2.00 |
| 680/6.3 | £3.00 |

*Wet tantalum*

| | |
|---|---|
| 220/25 axial | £3.20 |
| Castanet button 140/30, 470/3 | £3.20 |
| Hughes 540/10 | £3.20 |

*Bipolar*

| | |
|---|---|
| Hermetic bipolar tantalum 16/35 | £3.20 |
| Elcap axial 10/50 | £0.50 |
| Generic radial 100/16, 470/35, 100/ 63, 22/35, 4.7/35, 220/16 | £0.50 |

*Philips solid-aluminium (axial) 121/123*

| | |
|---|---|
| 47/16 | £1.00 |
| 330/6.3 | £2.00 |
| 100/35 | £4.00 |

*Philips Pearl 122 series (radial)*

| | |
|---|---|
| 10/16 | £0.30 |

*Silvered mica (radial)*

| | |
|---|---|
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*Trimmer capacitors*

| | |
|---|---|
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| Vishay plastic-film 5.5-45pF | £1.00 |
| Vishay plastic-film 5-80pF | £1.20 |
| Vishay plastic-film 10-250pF | £2.00 |
| Mica 1-12pF, 2-40pF | £1.00 |

## Audio transformers and inductors

*Eagle transformers*

| | |
|---|---|
| LT44, LT722 driver, LT700, LT723 500Ω output | £2.50 |
| LT30 500mW output | £3.50 |
| 5:1 interstage (fuzz box) | £1.50 |

*Repanco*

| | |
|---|---|
| T/T3 splitter transformer | £4.00 |
| CH2 5mH RFC | £2.00 |

*Balanced output transformer*

| | |
|---|---|
| Vigortronix 600Ω VTX-101-007 | £10 |
| Vigortronix 600Ω VTX-101-3001 | £10 |
| Vigortronix 600Ω VTX-101-3002 | £15 |
| Gardners 150Ω | £10 |
| Reclaimed BBC LL74/MPC nickel core 600Ω | £12 |
| **Low-level 1:1** 600Ω (modem) | £2 |
| Reclaimed mic input transformer | £15 |

*Inductors*

| | |
|---|---|
| 82µH, 4.7mH, 100µH, 270µH, 10µH, 14µH (low Z) | £0.50 |
| 4-inch ferrite rod with MW and LW windings | £3.00 |

## Special resistors

| | |
|---|---|
| Bourns wire-wound trimmer 5kΩ 3059 JM panel-mount | £2.00 |
| Thermistor RA53, R13 | £4.00 |
| A13 | £2.00 |
| Thermistor CZ1, CZ6 | £1.50 |
| Holco H2 2.2MΩ 1W, 1% | £1.00 |
| Welwyn 1GΩ 2W | £1.00 |
| 5k Bourns 3321H cermet trimmer | £0.50 |

## Potentiometers

*Bourns 81*

| | |
|---|---|
| 25kΩ lin cermet | £2.00 |
| 25kΩ lin, 5kΩ lin cond. plastic | £3.00 |

*Bourns 91*

| | |
|---|---|
| 10k dual-gang lin or log | £10.00 |

*Plessey moulded-track*

| | |
|---|---|
| 5kΩ log with switch | £3.00 |
| 50kΩ A/log | £3.00 |
| Mil 250kΩ lin dual | £5.00 |

*Alpha*

| | |
|---|---|
| 16mm 4.7kΩ A/log | £0.80 |
| 220kΩ A/log | £0.80 |
| 10kΩ lin centre-detent dual-gang | £1.50 |

*Allen Bradley J series/Honeywell*

| | |
|---|---|
| 10kΩ lin 1 million cycles | £5.00 |
| Blore Edwards AB 45 dual 5kΩ A/log with switch | £3.50 |
| Alps RK9 dual-gang 5kΩ RD law | £4.00 |
| Alps 50k log tapped motor. stereo | £5.00 |
| BI P260 500kΩ log conductive plastic 1 million cycles | £2.00 |
| Colvern wire-wound 100kΩ or 50kΩ dual-gang 3W | £5.00 |
| Mil Spec hermetic 10Ω | £8.00 |

## Miscellaneous

*PCBs*

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|---|---|
| **Pre-driver (*PE*, Jan26)** | **£3.00** |
| **Pre-driver jack bd (*PE*, Jan26)** | **£1.50** |
| **Fuzz Box (*PE*, Nov25)** | £3.00 |
| **Fuzz Box jack bd (*PE*, Dec25)** | £1.50 |
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| Mosfet Buffer (*PE*, Oct 2025) | £2.00 |
| Dual Mosfet Buffer (*PE*, Oct25) | £2.00 |
| Cascode Amp (*PE*, Sep 2025) | £2.00 |
| Cascode PS (*PE*, Sep 2025) | £2.00 |
| Pocket Theremin (*EPE*, 1996) | £2.00 |
| Elysian Theremin (*EPE*, 1996) | £6.00 |
| Synth VCF, VCO (*EPE*, 2017) | £3.00 |
| 48V PSU (*EPE*, 2019) | £3.00 |
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*Relays*

| | |
|---|---|
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*Switches Blore Edwards/Plessey*

| | |
|---|---|
| MH1 4-pole 3-way | £3.00 |
| Elma 01 2-pole 6-way | £6.00 |
| 3PDT guitar pedal stomp switch | £2.50 |

*Valves*

| | |
|---|---|
| EC88, PCL84 | £3.00 |

**Contact**

Jake Rothman
Manchester House
Market Street, Craven Arms
Shropshire SY7 9NN
(visit by appointment)
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# AUDIO OUT

By Jake Rothman

## The Pre-driver guitar overdrive pedal

We'll now move on to the guitarist's next favourite pedal after fuzz, which is overdrive. This is a preamp/tone control circuit used to overload/drive the guitar amp. This works most effectively if the guitar amplifier is a valve-based design. Typically, the first ECC83 triode stage in the amp can be overdriven, which the guitar alone can't do.

The amplifier's master volume control can then be turned down, giving some of the sound of an overdriven valve amplifier. This avoids the ear-splitting levels that occur if the output stage is driven into overload.

However, most guitarists and sound engineers still want the full AC30 amp overdrive sound from a Celestion G12 speaker cone breaking up along with push-pull EL84 output valve slope distortion, so hearing damage it is. When recording, I used to confine the guitarist and his amp to an isolation booth, but I still suffered hearing damage going in there to find the optimum mic position.

Most overdrive pedals are fairly clean in themselves, with typically 0.2% total harmonic distortion (THD), although some do generate a bit of their own fuzz at high outputs.

Guitar pedals have a unique culture of funny names, some rather rude. The Klon and Tube Screamers (variations on the Ibanez original) are popular. Two designs I have been involved with have been the Colorsound Overdriver (Photo 1) and the Powerboost (Photo 2), from which this circuit is derived.

To avoid infringing anyone's name, especially litigious Americans, I decided to call this pedal the Pre-driver. Yes, there is a pedal called something similar, but I used a smug hyphen, so I claim it.

### Circuit structure

Fig.1 shows the basic system: an input buffer, preamp stage and tone control. There may also be a master volume on the output, allowing the pedal to distort if needed. The tone control is occasionally just a treble boost, as in the Electro-Harmonix Soul Food, but usually at least bass and treble controls are provided.

There was a variation of the Electro-Harmonix pedal by JHS with a "meat" modification (to offend vegans?), but it did sound "meatier".

### More gain, more pain

Most guitar pedals work at a nominal signal voltage of -10dBu or 0.7V peak-to-peak. The "u" means "not referenced to any impedance"; it's just the voltage level. This is because the high impedances involved don't result in any significant power.

Most studio equipment and devices like CD players use 'line level', 0dBu, which is 2.2V peak-to-peak. Professional equipment often uses +4dBu, which is 3.5V peak-to-peak. It is often stated using the old telephone standard, dBm, referenced to 600Ω, but that is rarely the case in reality.

The difference between guitar and studio levels often results in clipping if guitar pedals are used as studio effects. In this design,

INSTRUMENT
AMPLIFIER
COLORSOUND
LONDON·ENGL
TREBLE
BASS
POWER BOOST
PRE-SE
BUTTON
HIT IT!

Photo 2: the Colorsound Powerboost is an overdrive pedal with an added JFET input buffer.

INSTRUMENT
AMPLIFIER
COLORSOUND
LONDON·ENGL
TREBLE
BASS
DRIVE
OVERDRIVER
PRE-SE
BUTTON

Photo 1: the Colorsound Overdriver embodies all the standard functions of a good guitar preamplifier.

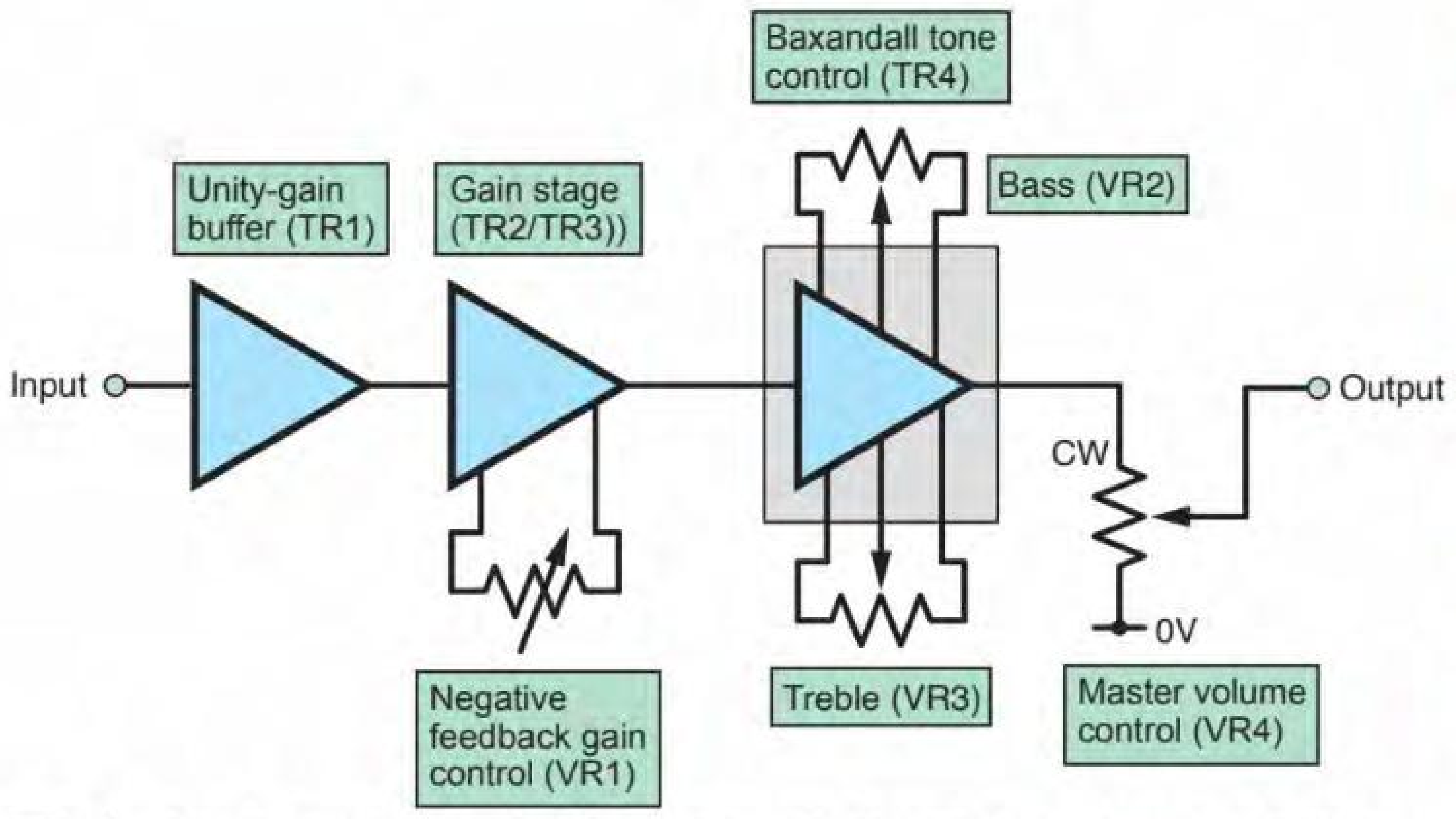

Baxandall tone control (TR4)
Unity-gain buffer (TR1)
Gain stage (TR2/TR3))
Bass (VR2)
Input
Output
CW
0V
Negative feedback gain control (VR1)
Treble (VR3)
Master volume control (VR4)

Fig.1: the basic block diagram of most overdrive pedals.

the gain of the preamp varies from +17dB (7×) to +47dB (200×), while the tone control is ×1 or unity gain.

When Gary Hurst designed the Colorsound Powerboost, he based it on the 1969 Mullard "Hi-Fi" preamp in their *Transistor and Radio Circuits* book (page 142 in the 1972 2nd edition) and its many variants published in *Practical Wireless*.

This had a two-stage DC-coupled gain stage and a single-stage Baxandall tone control running on 18V. It used the silicon planar, metal-cased Mullard BC109 transistors, which were then the latest thing.

My take on the later Dick Denny production circuit, shown in Fig.2, is very similar but with the resistor values reduced to reflect the lower supply voltage. Colorsound Overdrivers made from 1972 used a modern plastic-encapsulated equivalent of the BC109, the Texas Instruments BC184L (which had a central collector pin).

We'll use the latest incarnation of the BC109, the BC549, made by Philips and later by Fairchild. They are still the go-to low-noise, low-voltage, high-$h_{FE}$ small-signal audio transistor, and they work out about 2p each if you buy a bag of 100.

## More pot problems

A major modification required to the Mullard circuit was to make the gain variable. This was achieved by adding a pot, VR1, to the lower arm of the feedback network. Originally this was a 4.7kΩ linear pot, which gave an abrupt boost in gain at the end of its rotation. That made it difficult to use and could result in uncontrolled acoustic feedback if not adjusted carefully.

Later on, this was changed to a 2.2kΩ anti-log pot ("C" taper), needed because the gain goes up as the resistance gets smaller, the opposite to a normal volume pot. Unfortunately, anti-log or reverse-taper pots are hard to get and are normally specially made (by Omeg in the case of Colorsound; see Photo 3).

Tayda makes Alpha antilog pots, but these are difficult to mount. A solution to this is to put the pot in the upper arm of the feedback network, allowing a normal log pot to be used. This can't be done here, since it would upset the DC bias conditions. I developed a circuit using a normal log pot in Fig.3, but it had less gain.

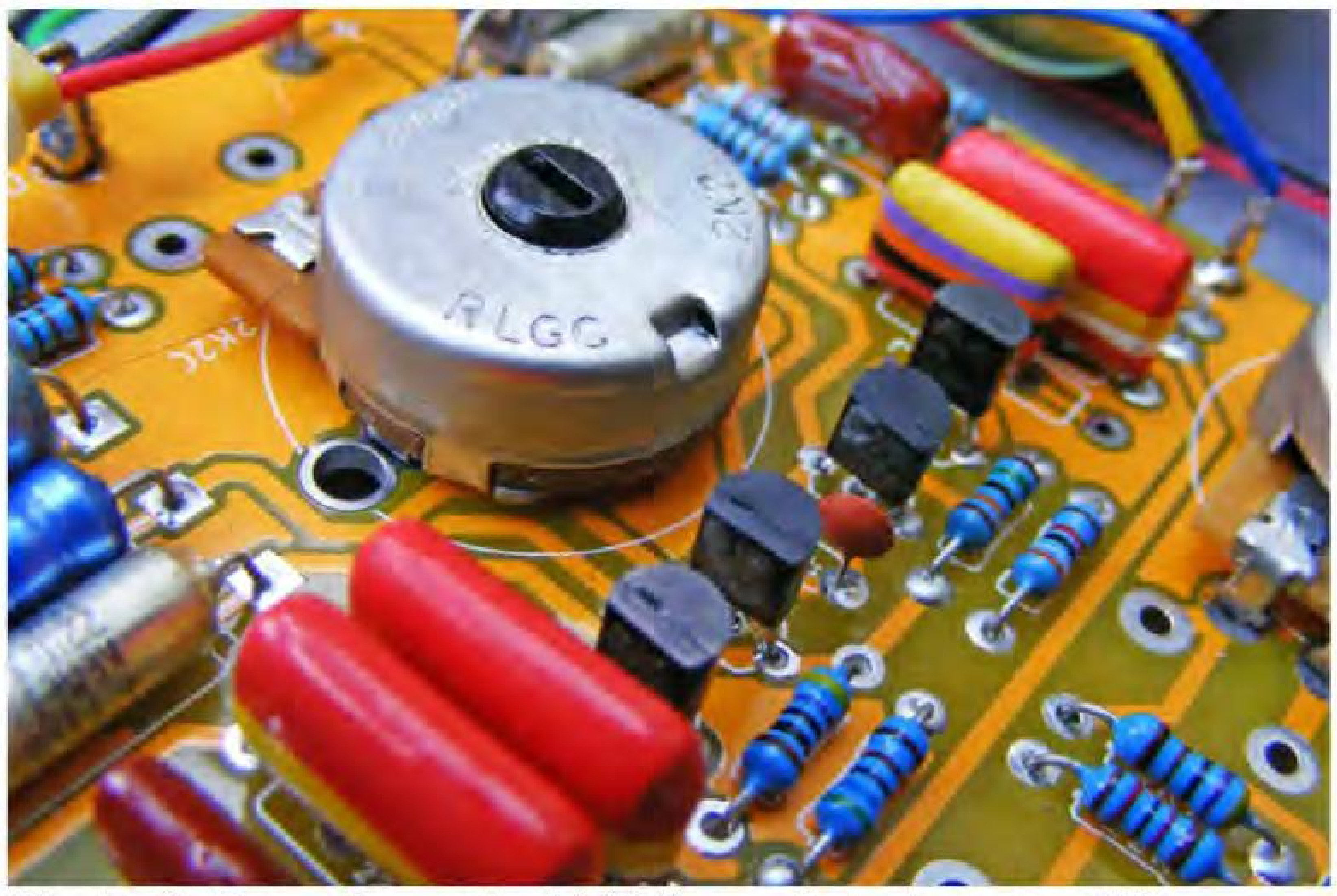


Photo 3: the gain control (VR1) needs to be an anti-log taper pot, written R-LOG on this Omeg example.

Diodes D1 and D2 prevent DC bias shift when overdriven. The input impedance is lower than the original design, so the input buffer needs a 2.2kΩ source resistor.

Another problem with gain control pots is getting the gain to go low enough. This can't be done unless

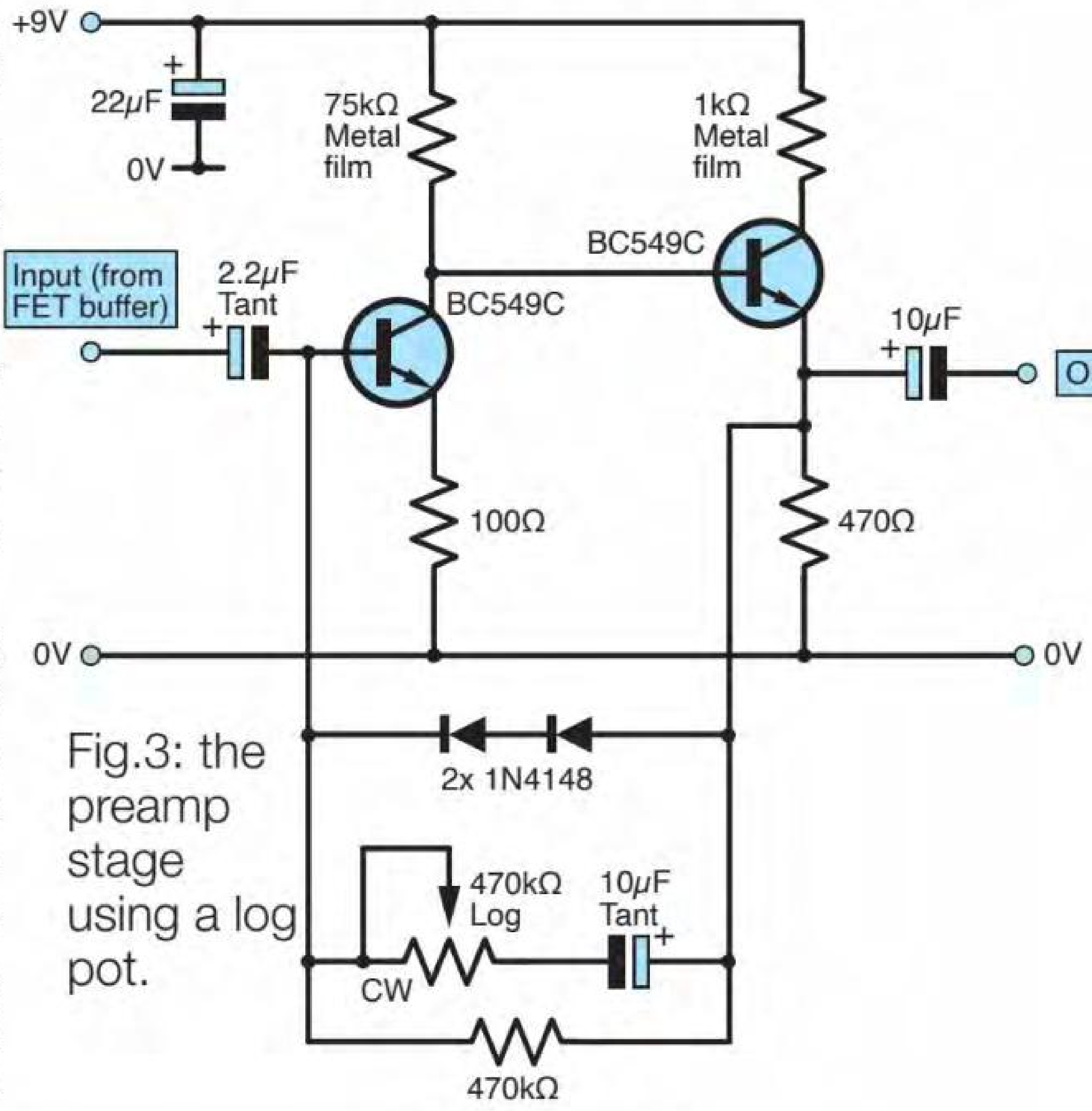

+9V
22µF
0V
75kΩ Metal film
1kΩ Metal film
BC549C
BC549C
Input (from FET buffer)
2.2µF Tant
10µF
Output
100Ω
470Ω
0V
0V
2x 1N4148
470kΩ Log
10µF Tant
CW
470kΩ

Fig.3: the preamp stage using a log pot.

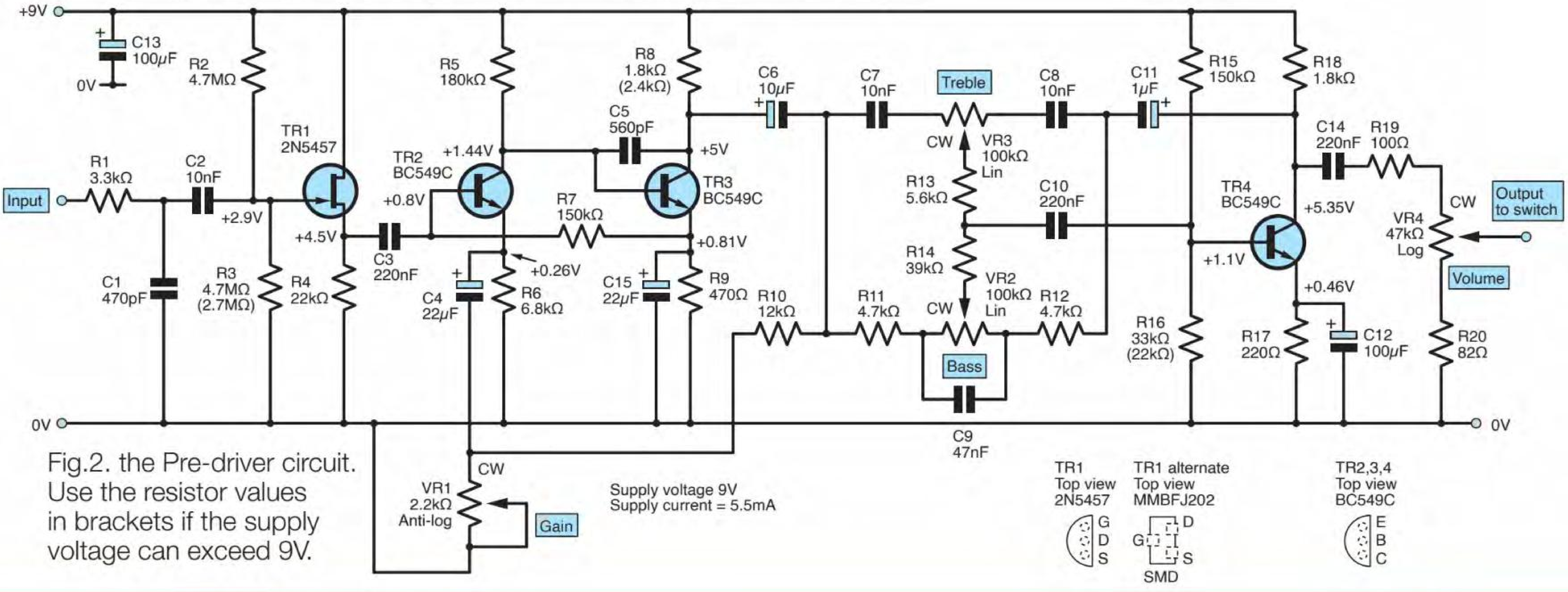

+9V
C13 100µF
0V
R2 4.7MΩ
R5 180kΩ
R8 1.8kΩ (2.4kΩ)
C5 560pF
C6 10µF
C7 10nF
Treble
C8 10nF
C11 1µF
R15 150kΩ
R18 1.8kΩ
TR1 2N5457
TR2 BC549C
+1.44V
+5V
TR3 BC549C
CW
VR3 100kΩ Lin
R13 5.6kΩ
C10 220nF
C14 220nF
R19 100Ω
R1 3.3kΩ
C2 10nF
Input
+2.9V
+0.8V
+4.5V
R7 150kΩ
+0.81V
TR4 BC549C
+5.35V
CW
Output to switch
VR4 47kΩ Log
+1.1V
Volume
C1 470pF
R3 4.7MΩ (2.7MΩ)
R4 22kΩ
C3 220nF
+0.26V
C4 22µF
R6 6.8kΩ
C15 22µF
R9 470Ω
R10 12kΩ
R11 4.7kΩ
R14 39kΩ
VR2 100kΩ Lin
CW
R12 4.7kΩ
Bass
+0.46V
R16 33kΩ (22kΩ)
R17 220Ω
C12 100µF
R20 82Ω
0V
0V
C9 47nF
CW
VR1 2.2kΩ Anti-log
Gain
Supply voltage 9V
Supply current = 5.5mA
TR1 Top view 2N5457
G
D
S
TR1 alternate Top view MMBFJ202
D
G
S
SMD
TR2,3,4 Top view BC549C
E
B
C

Fig.2. the Pre-driver circuit. Use the resistor values in brackets if the supply voltage can exceed 9V.

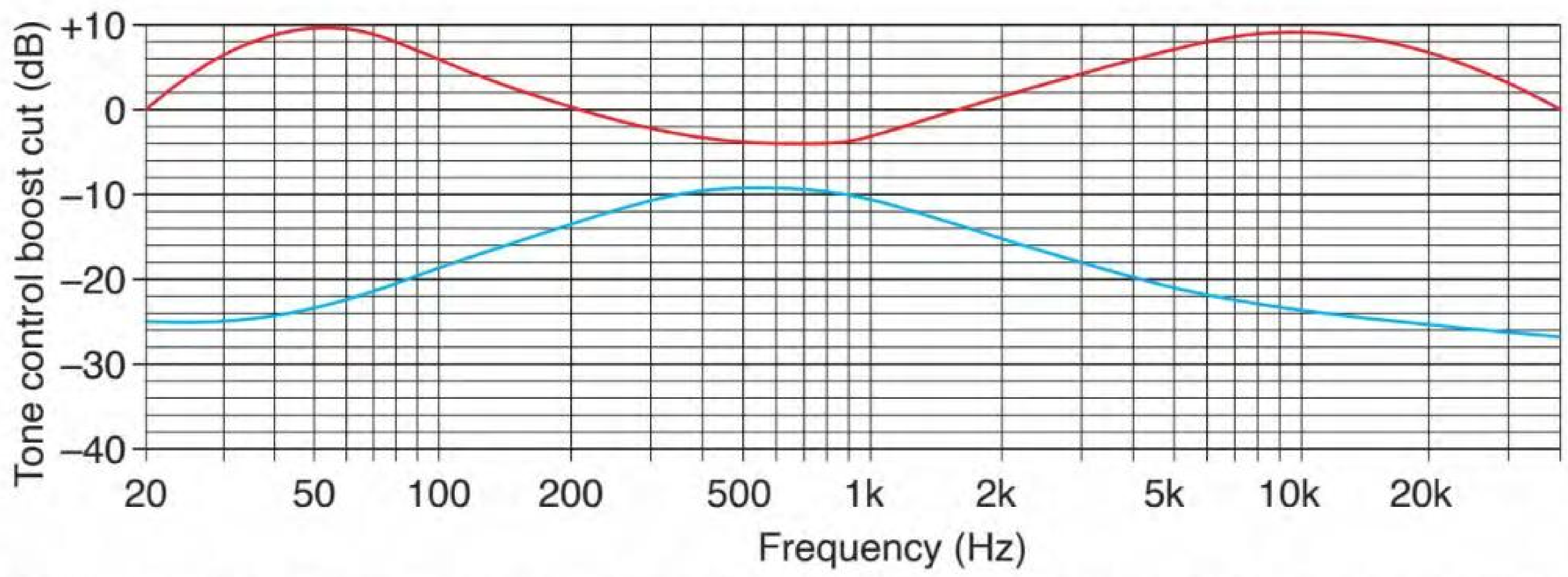


Fig.4: the tone control curves. Note the scale is compressed by the use of the 50dB scale. The boost and cut is ±14dB at 50Hz and 10kHz.

the CW terminal of the 2.2kΩ log pot is switched out to give full negative feedback. A gain of 4.4dB (1.66×) is then obtained. You try getting an anti-log pot with a switch!

I do have some dual-gang 5kΩ anti-logs with switches that I had specially made for mic preamps. With the two gangs connected in parallel, they work well here, but their body is 40mm deep, which means that they won't fit in a low-profile pedal box.

They would have to go on the side; it's simpler to put in a separate switch.

At the moment, you have to cut the pot tag, which is a bit messy. I'll put a proper switch connection on the next batch of boards.

### Input buffer

I think guitars sound better feeding high-impedance JFET inputs, so I added a JFET input buffer to the circuit. The Colorsound Overdriver is actually the original Powerboost, having no buffer, while the current Powerboost has one.

The "Aussie" Powerboost variant, made for Pink Floyd tribute bands, generates a spike at the verge of clipping achieved by leaving off C15.

This 'upsets' the distorting guitar amplifier in an interesting way. This helps achieve that David Gilmour (of Pink Floyd fame) sound.

Queen guitarist Brian May is famous for using a simple germanium treble booster to drive a tiny transistor amplifier built by bassist John Deacon in 1972. The half-watt amp, based on a Class-AB germanium output stage and fitted with a six-inch Elac speaker, produced a surprisingly rich, saturated tone when overdriven.

May used it extensively for layered guitar harmonies on Queen recordings. The unit became known as the Deacy Amp and it remains an integral part of May's distinctive sound (I described the Deacy Amp in some detail in June 2015).

### Volume control

A master volume control was added to the Powerboost and Overdriver around in 1990, mounted on the side of the box.

This is done here as well, using VR4. It enables it to be used as a fuzz box to a degree, but it is not as good as a dedicated unit because the high negative feedback used means the onset of distortion is abrupt.

The Pre-driver works well driving power amps directly to make a complete guitar amp.

### Tone control

The Baxandall tone control stage is completely conventional, although the 100kΩ pots would be considered too high in value today in a hi-fi preamp because of the high Johnson noise generated.

The value used in the original circuit was 220kΩ. I found that using more readily obtainable 100kΩ potentiometers made no difference. The tone control curves are shown in Fig.4.

A single-transistor stage would also be frowned upon in hi-fi these days because the THD at full boost is around 1%. However, for electric guitarists, it doesn't matter because the hum and noise of the high-impedance magnetic pickups completely dominates the noise floor, and distortion is actually a goal.

I suspect some guitarists even like 'resistor grunge', with some pedal makers using original carbon-composition resistors. I spend a lot of my time removing them from old audio equipment.

One advantage of using high-value resistors, of course, is low current consumption. The Pre-driver draws just 5.6mA from the battery, leading to a long battery life, which is important for those extended jam sessions!

### Power supply

It's standard for most guitar pedals to run on a 9V 'PP3 style' block battery, which is the minimum voltage to avoid signal clipping at standard guitar signal output levels.

In this case, clipping occurs with a 5V peak-to-peak signal level with the tone controls flat, the gain set to its minimum and the volume control at its maximum.

Increase the supply voltage for a more effective overdrive and less chance of hard transistor clipping. A higher voltage can be obtained by clipping two PP3s together in series, or by using an external power supply, like a plugpack (eg, 12V DC or 15V DC). Colorsound offers an 18V

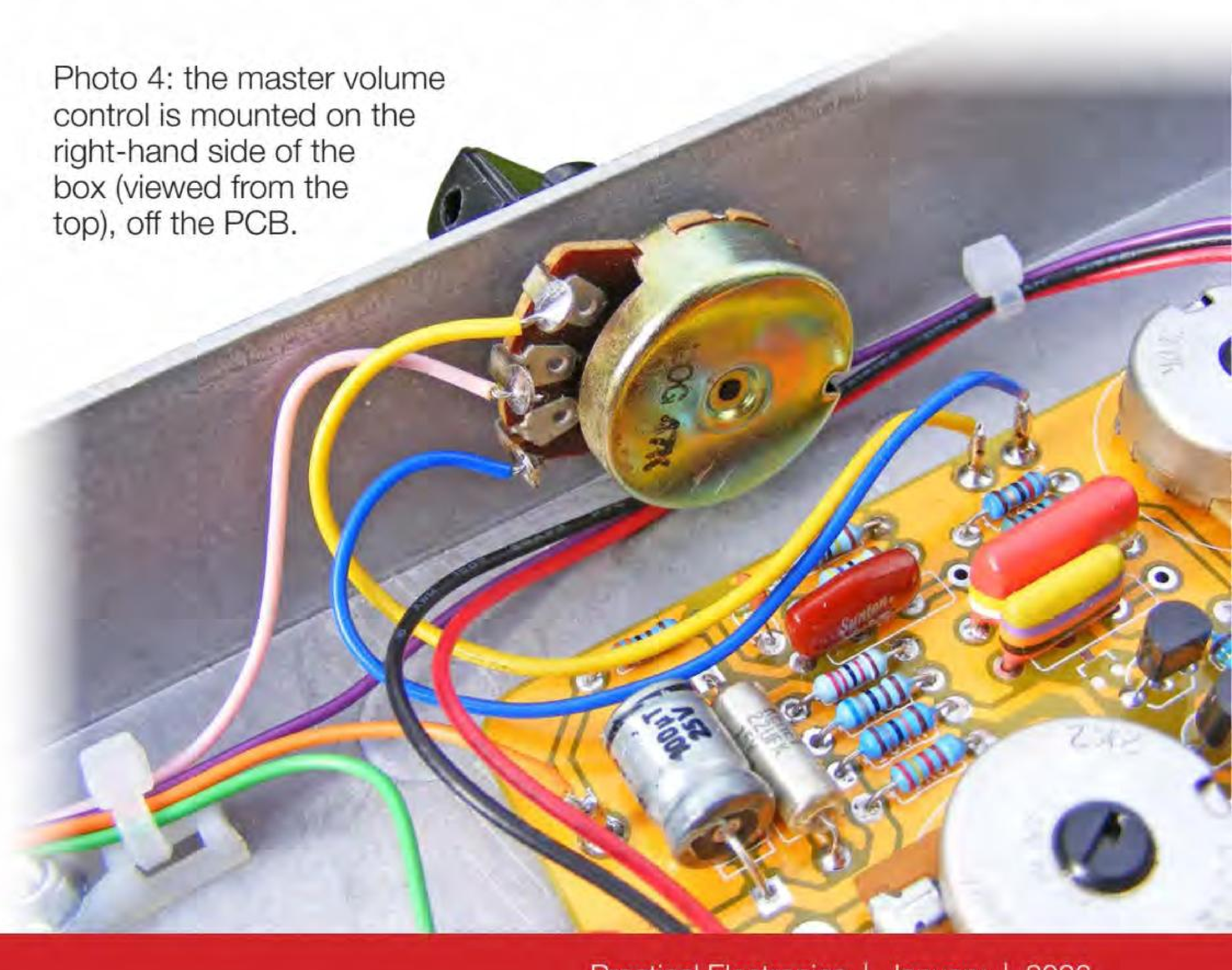


Photo 4: the master volume control is mounted on the right-hand side of the box (viewed from the top), off the PCB.

original version of the Powerboost built by Stu Castledine for Macari's.

Some pedals use an up-converter, but this can cause switching noise break through and it will also flatten a PP3 battery much more quickly (a higher output voltage implies a higher input current; there's no such thing as a free lunch!).

This design is re-biased so that it can be used at either voltage. The original 9V values are on the PCB, with the values for a higher supply voltage (up to 18V) shown in brackets on the circuit diagram in Fig.2.

The current consumption at 18V increases to 13mA.

### Building it

The only extra bit to the construction compared to the Transmanium fuzz box, described over the last two issues, is the master volume control potentiometer mounted on the side of the box, as shown in Photo 4. We'll get to assembly of the main circuit board soon, but first…

### Drilling

The best way to drill the PCB mounting and pot holes is to mark their positions on the case using the PCB as a template. I find I'm more prone to errors when measuring them with a ruler. I have also prepared a drilling diagram, available at **https://pemag.au/link/ac9b**

I use 10mm holes for the potentiometers rather than the 'correct' ⅜-inch/9.5mm to give a bit of clearance/wiggle room. To prevent the drill bit wandering, make an indentation with a centre punch (or, in a pinch, a nail hit with a hammer).

It's always wise to use a deburring tool on the holes to remove sharp ridges (you can use a larger drill bit, twisted by hand, as long as it's sharp).

It is especially important to remove any paint on the edges of the holes to ensure good Earthing of the component bodies, since the case is grounded through the pot bushes from the PCB. Also, touching the knob of a floating pot can cause hum to be injected into the signal.

Even the 12.5mm switch and 10mm master volume holes must be clean to Earth their metal bodies. Marking out for the jack board is more difficult, especially the 8mm DC connector hole, which is 1.5mm higher than the jack holes (11mm).

### Jack board

I took this second pedal as an opportunity to redesign the jack board. By making the holes for the

## Parts List – Pre-driver pedal

1 main PCB, 84 × 74mm, coded AO-JAN26-1
1 jack board MkII, 72.5 × 25mm, coded AO-JAN26-2
1 Tayda 1590DD 190 × 121 × 37mm diecast aluminium case
1 PP3 battery snap/connector with 200mm-long wires **[Rapid 18-0093]**
1 battery mounting spring clip **[Rapid 18-3480 – Comfortable BH980]**
1 PCB-mounting 2.1mm ID barrel socket **[Rapid 20-0970 TruConnect, Tayda A-4118]**
2 stereo 6.3mm PCB-mounting jack sockets **[Tayda A-5079]**

**Hardware & wire**
6 80 × 3mm cable ties **[Rapid 04-0630]**
1 cable tie mounting clip/eyelet **[Rapid 04-0600]**
10 500mm lengths of 7/0.2 red, orange, yellow, green, blue, purple, pink, brown, grey & black equipment wire
3 Tayda A-7023 potentiometer shaft adaptors
3 ⅜-inch potentiometer nuts **[Mouser 594-511-3002, MPN Vishay ACCTR511-3002]**
3 M7 Alpha potentiometer nuts **[Tayda A-5037]**
3 M3/M4 potentiometer grub screws (if needed) **[Tayda A-7804 or A-7805]**
6 12.5mm rubber hole plugs **[Farnell 896603]**
6 M3 × 12mm countersunk head screws and hex nuts

**Semiconductors**
1 2N5457 N-channel JFET (TR1)
3 BC549C NPN bipolar transistors (TR2-TR4)
1 3mm orange 3mm high-efficiency LED
**[Tayda A-264, Rapid 72-8972 (preferred) or Kingbright L-7104SEC]**

**Capacitors**
1 100µF 25V electrolytic (C13)
1 100µF 6.3V electrolytic (C12)
2 22µF 10V electrolytic (C4*, C15)
1 10µF electrolytic (C6*)
1 1µF 10V electrolytic (C11*)
3 220nF polyester ±20% (C3, C10, C14)
1 47nF polyester ±20% (C9)
3 10nF polyester ±10% (C2, C7, C8)
1 560pF ceramic disc ±10% (C5)
1 470pF ceramic disc ±10% (C1)
* can be replaced with tantalum for lower pot rotational noise

**Potentiometers* (all ⅜in 10mm-long bush, ¼in shaft, >20mm diameter body)**
2 100kΩ or 220kΩ single-gang linear (VR2, VR3)
1 22kΩ or 47kΩ single-gang log (VR4)
1 4.7kΩ or 2.2kΩ single-gang anti-log (VR1)
4 pointer knobs to suit VR1-VR4 (four different colours if possible) [RS or Tayda]
* preferably Omeg (best) or Alpha (good) brand

**Resistors (all ¼W ±5% metal or carbon film)**
2 4.7MΩ (R2, R3)
1 180kΩ (R5)
2 150kΩ (R7, R15)
1 39kΩ (R14)
1 33kΩ (R16)
1 22kΩ (R4)
1 12kΩ (R10)
1 6.8kΩ (R6)
1 5.6kΩ (R13)
2 4.7kΩ (R11, R12)
1 3.3kΩ (R1)
2 1.8kΩ (R8, R18)
1 470Ω (R9)
1 220Ω (R17)
1 100Ω (R19)
1 82Ω (R20)

Substitutions for 18V operation:
R3: 4.7MΩ → 2.7MΩ
R8: 1.8kΩ → 2.4kΩ
R16: 33kΩ → 22kΩ

**All parts listed here are available individually from the AOShop (see page 49).**

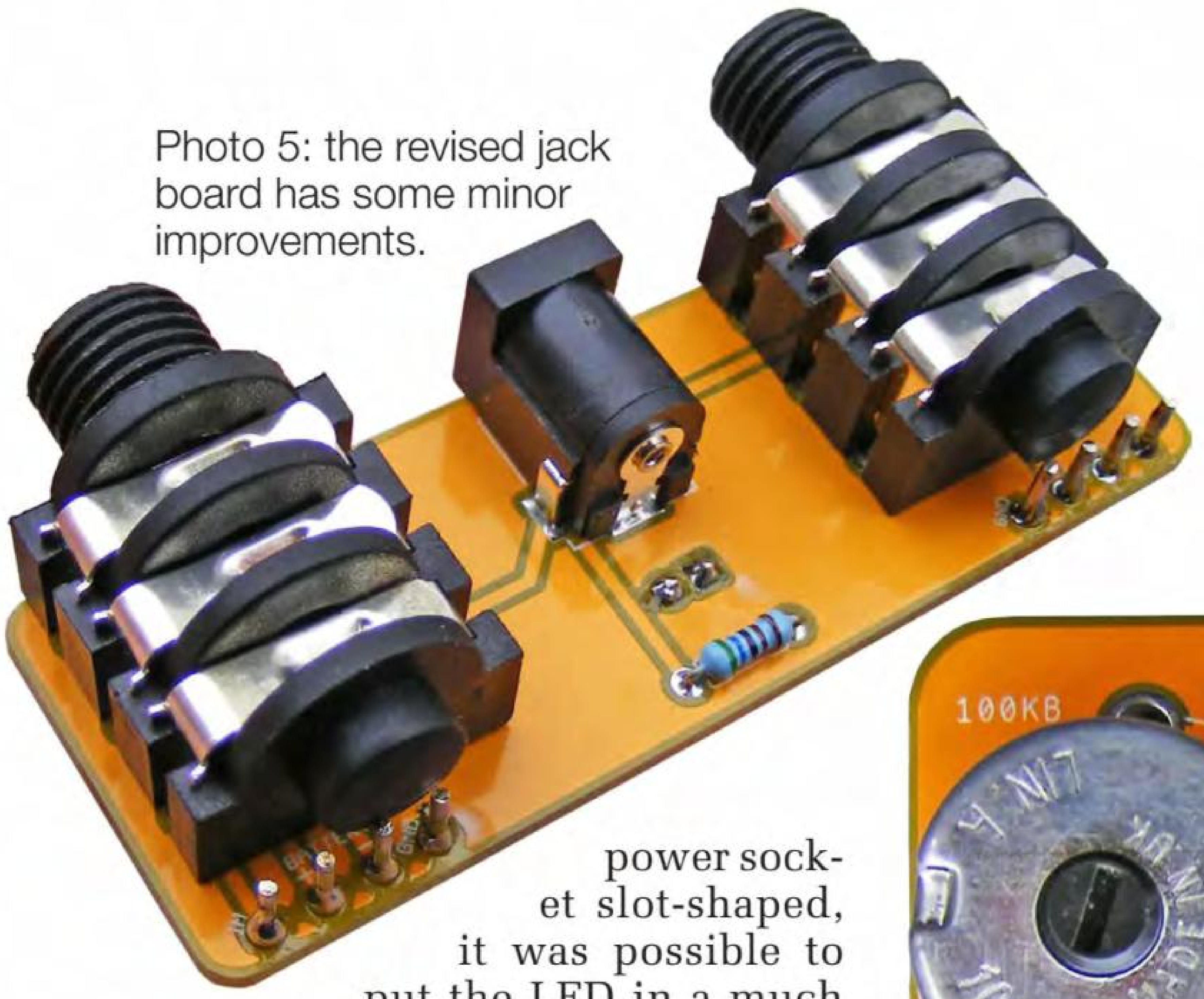
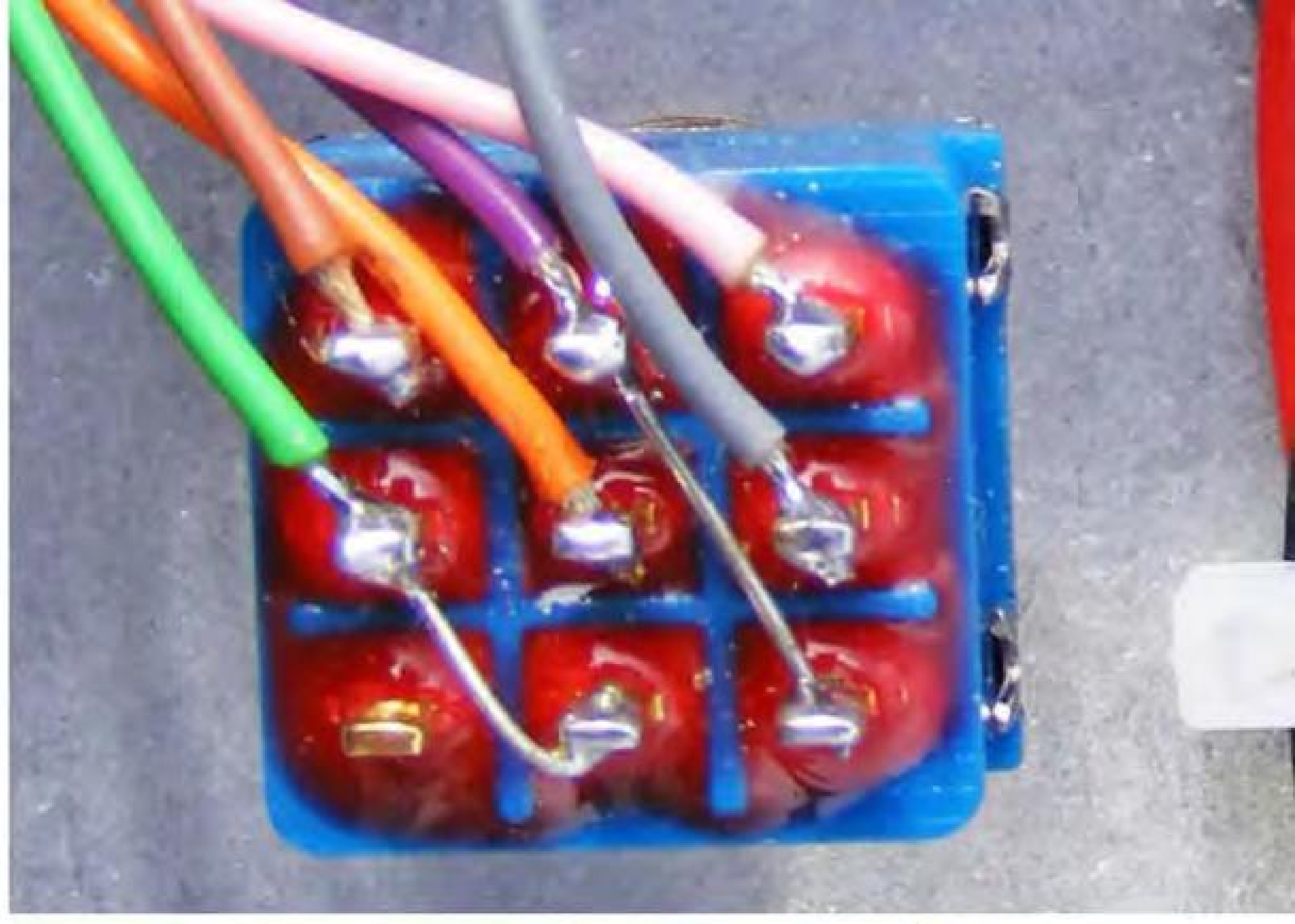


Photo 5: the revised jack board has some minor improvements.



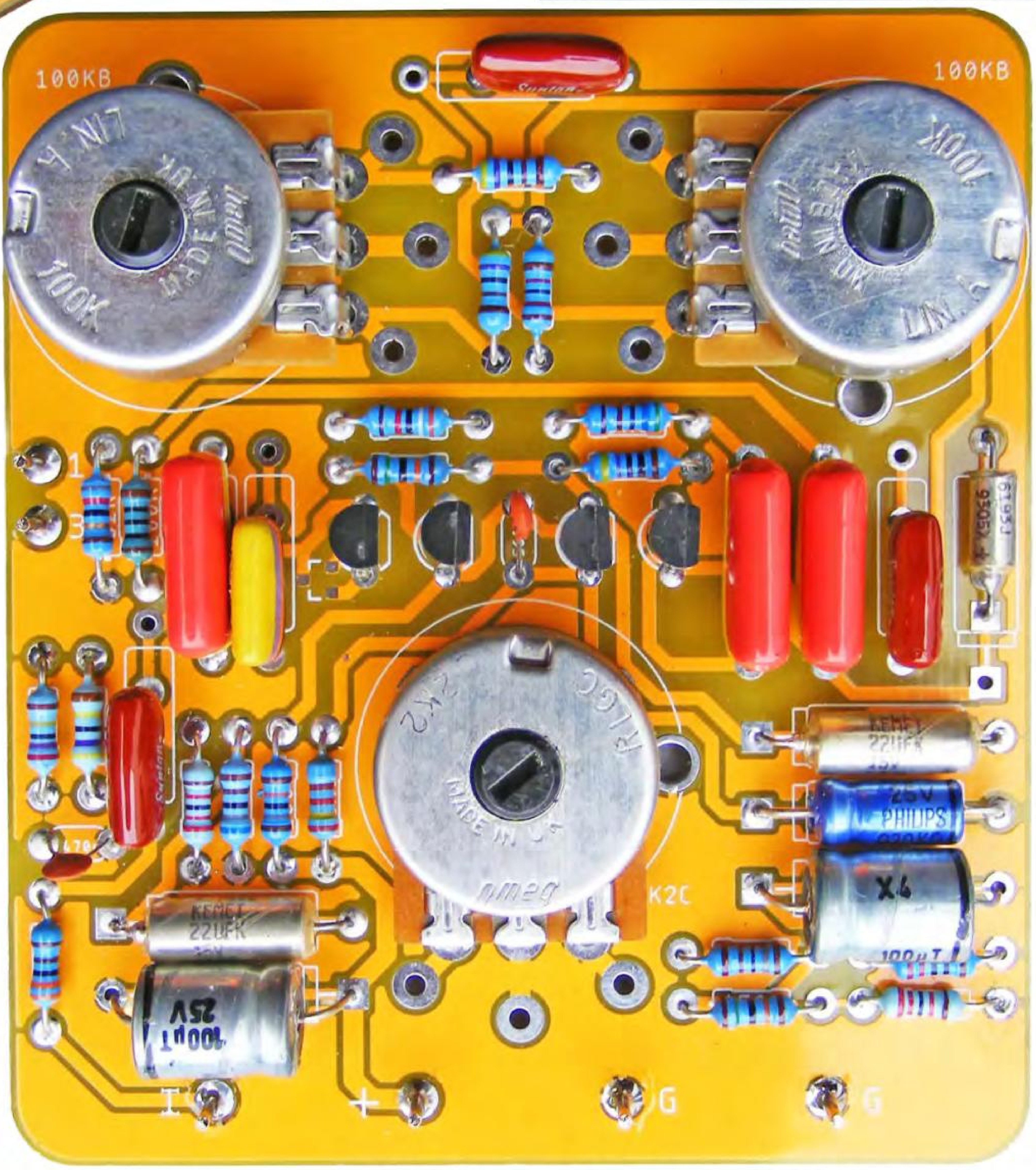

100KB
100KB
1
3
K2C
I
+
G
G

Photo 6: the main PCB, with reassuring symmetry and pretty C280 capacitors. This is an interesting example of where poor-quality components can ▼ boost sales!

power socket slot-shaped, it was possible to put the LED in a much better position, avoiding the need for wiring. This should be soldered after the board is mounted to prevent strain. I also reversed the input and output sockets, in line with guitar convention.

The new Mk II jack board overlay is shown in Fig.5 and is seen assembled in Photo 5. It is not essential to use the jack board; some constructors may prefer to hardwire it. Nothing's ever completely finished in electronics. I'm now doing a Mk III jack board because I want the battery wires on the left where the battery clip is. Right now, the 200mm battery connector is only just long enough. I'll show this board next time.

## Main PCB

The main PCB overlay is shown in Fig.6 and you can see what the assembled board looks like in Photo 6. An SMT/SMD JFET option for TR1 is provided, visible next to the yellow-topped capacitor. For a while, discrete leaded JFETs became very expensive. However, I noticed that the situation has abated somewhat; for example, BF256s now cost just 16p at Rapid. The SMT devices are still cheaper, though.

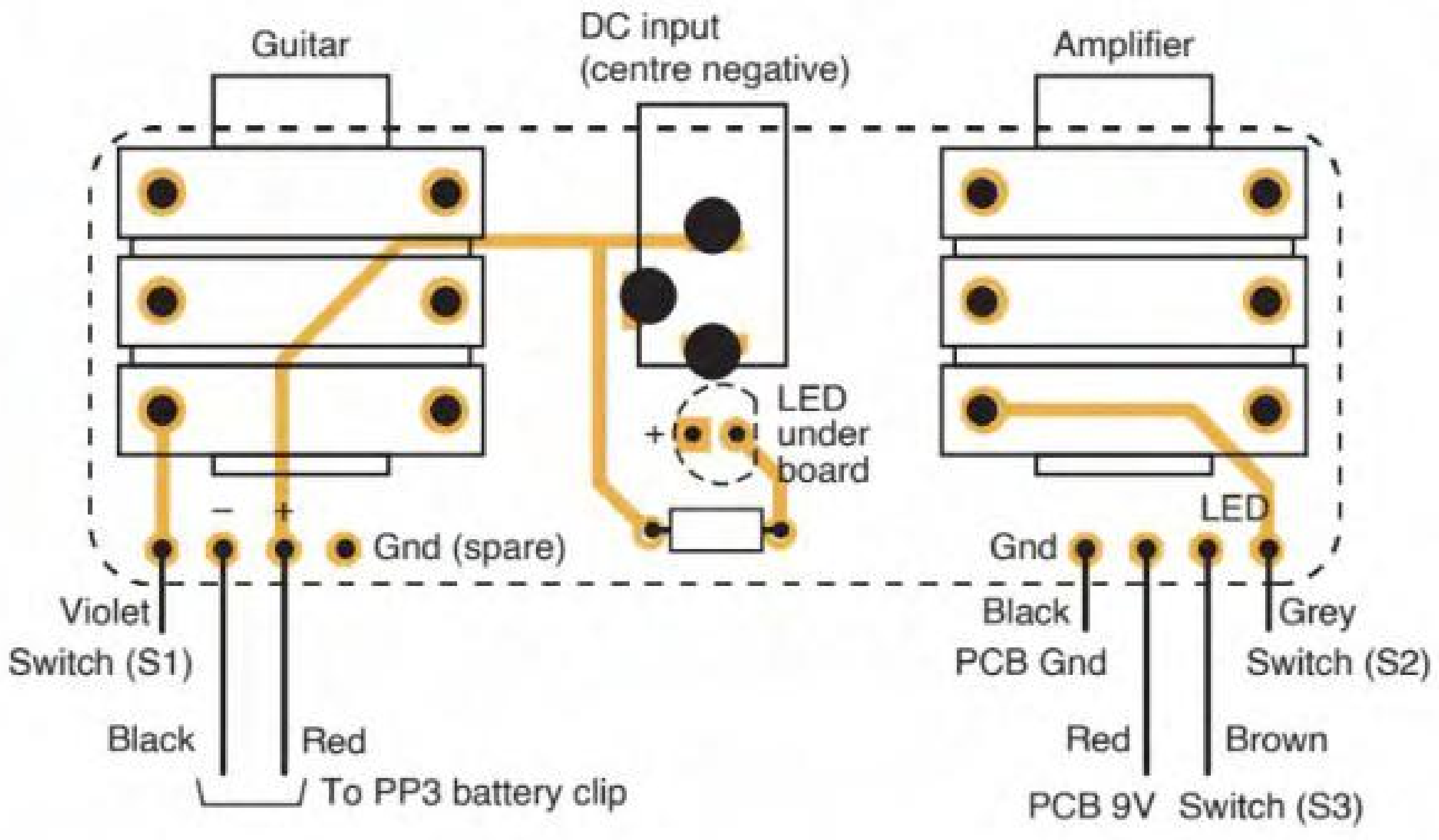

Guitar
DC input
(centre negative)
Amplifier
LED
under
board
+
–
+
Gnd (spare)
Violet
Switch (S1)
Black
Red
To PP3 battery clip
Gnd
LED
Black
PCB Gnd
Grey
Switch (S2)
Red
PCB 9V
Brown
Switch (S3)

Fig.5: the new jack PCB. Some early boards may be marked for positive Earth. Ignore that; follow this overlay.

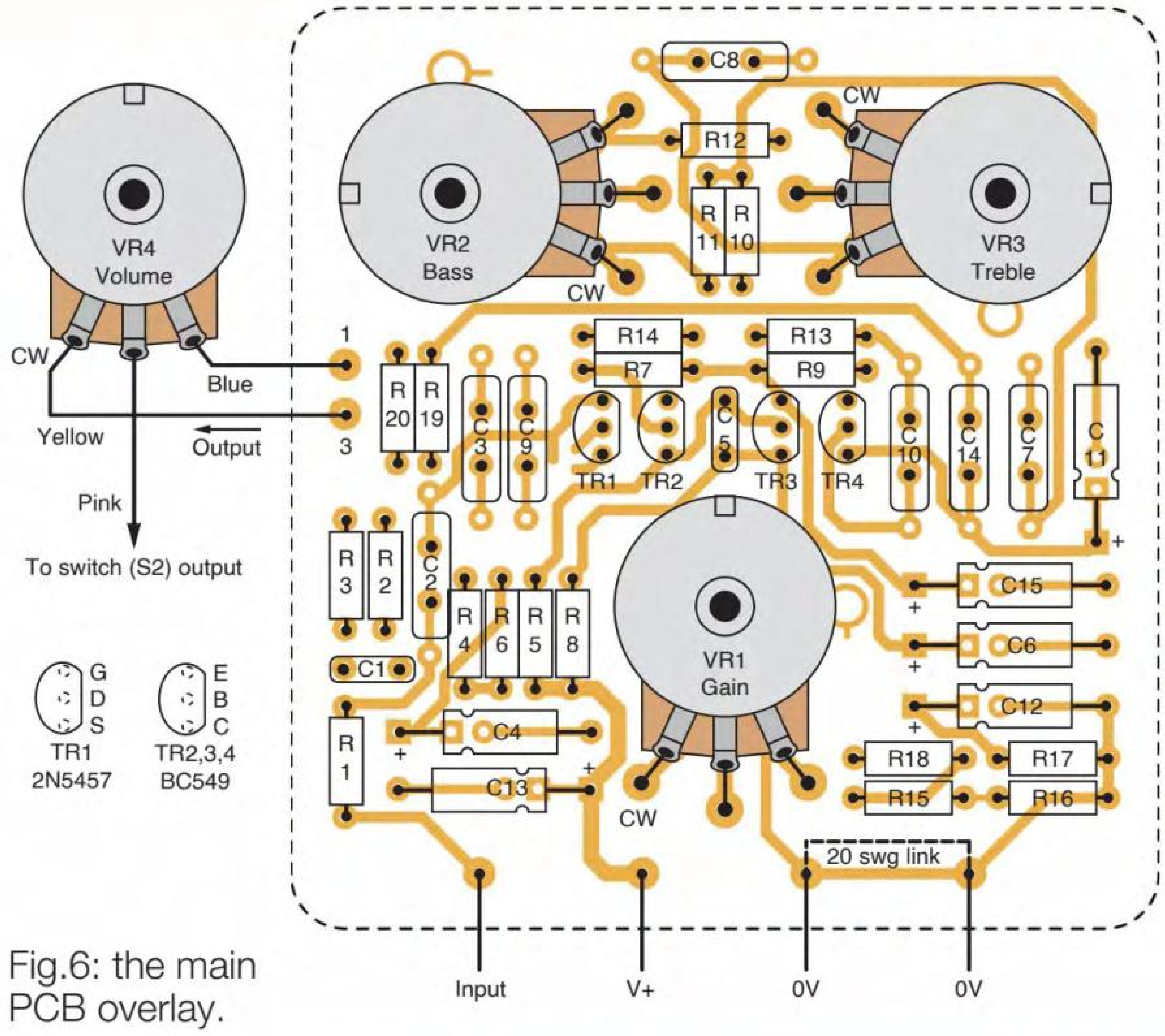

C8
R12
CW
R 11
R 10
VR2
Bass
VR3
Treble
CW
VR4
Volume
CW
Blue
Yellow
Output
Pink
To switch (S2) output
1
3
R14
R13
R7
R9
R 20
R 19
C3
C9
TR1
TR2
C5
TR3
TR4
C10
C14
C7
C11
+
R3
R2
C2
R4
R6
R5
R8
VR1
Gain
C15
C6
C12
+
C1
C4
C13
R1
+
R18
R17
R15
R16
CW
20 swg link
Input
V+
0V
0V
G
D
S
TR1
2N5457
E
B
C
TR2,3,4
BC549

Fig.6: the main PCB overlay.

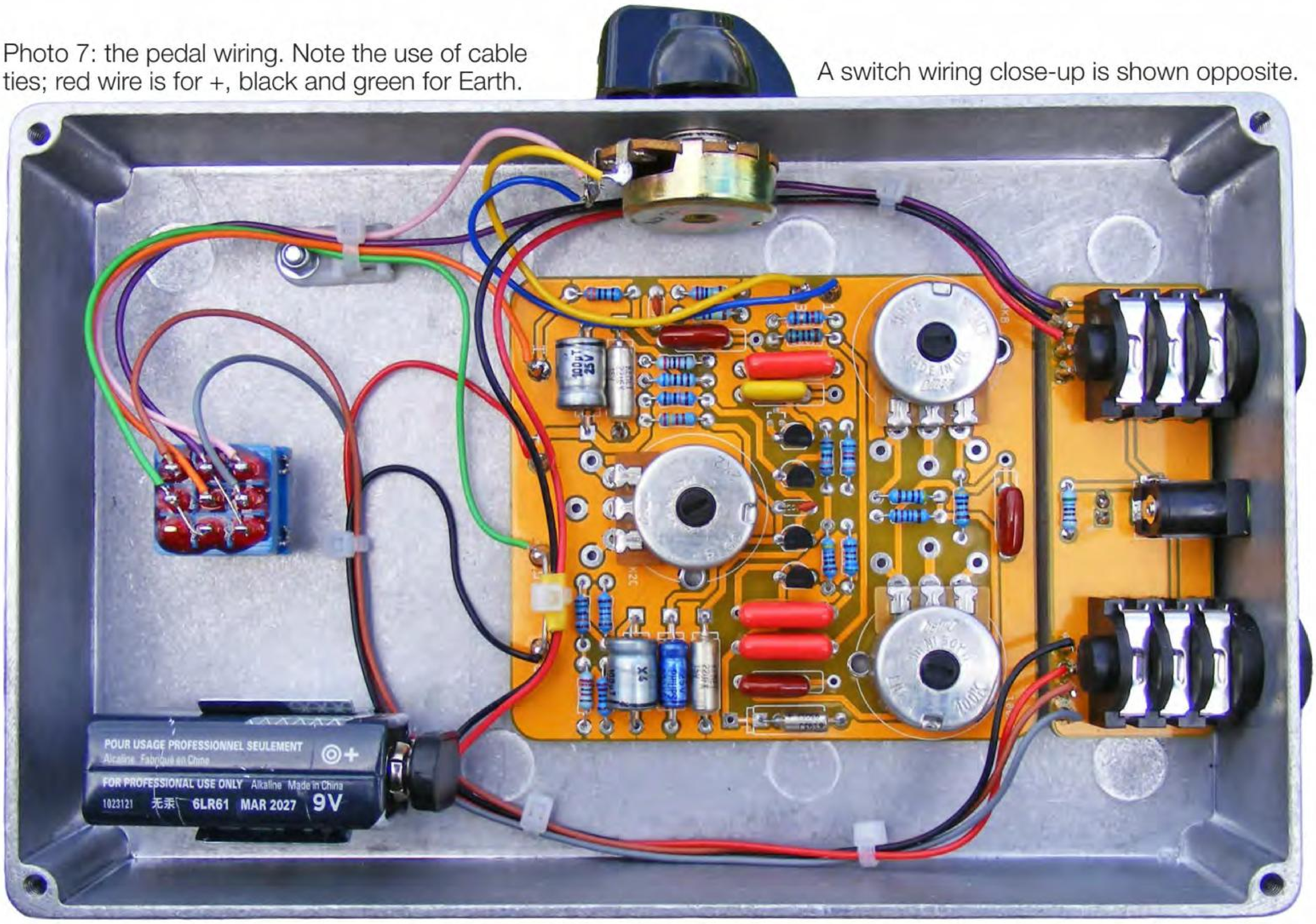


Photo 7: the pedal wiring. Note the use of cable ties; red wire is for +, black and green for Earth.

A switch wiring close-up is shown opposite.

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6LR61 MAR 2027
9V

I used the J201 from Tayda, which has lower noise (the J202 is also suitable). Although source-followers are fairly insensitive to different JFETs (because of the 100% negative feedback), it still might be necessary to adjust R2 or R4 for maximum headroom. It is worth checking if you use something very different from the ubiquitous 2N5457.

Be careful with the pin-out for the gate. Sometimes it's the middle, upper or lower pin. Since JFETs are symmetrical at audio frequencies, the drain and source pins can be swapped, making substitution easier.

The assembly procedure is virtually the same as the fuzz pedal given last month, except that the component values and positions are different, of course. Pay attention to the orientations of the transistors and axial electrolytic capacitors; they must all be orientated as shown in Fig.6 for the circuit to work properly.

## Wiring up

The need for screened leads can be avoided if the purple input lead is well tucked in along the corner of the metal box. Instability is avoided by having the output lead run up the other side. The inverting nature of the circuit also helps. This is all shown in Photo 7.

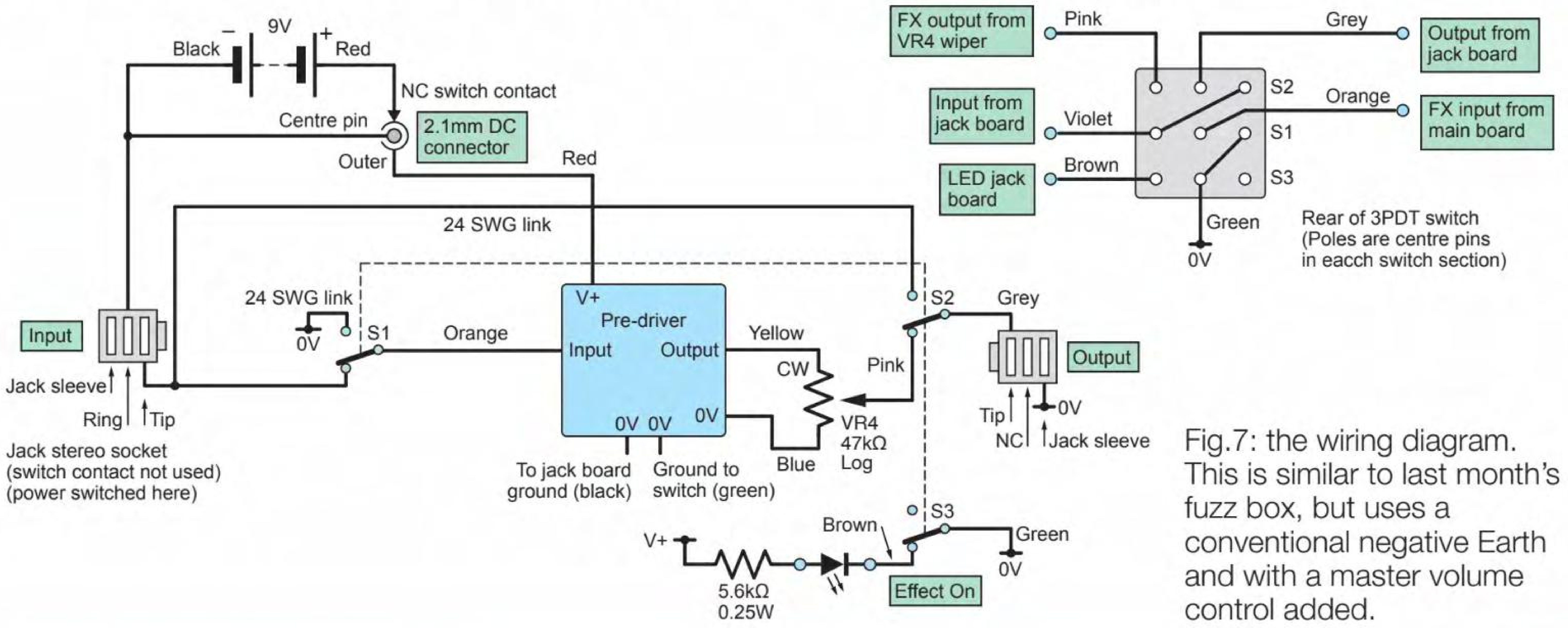
A little trick here is using a loop of thick (20SWG; 0.9-1mm diameter) wire between the two Earth connections on the lower-left of the PCB to act as a tie point for the battery leads.

## Finishing up

Unpainted diecast boxes labelled with Dymo tape are functional but have a rather austere look to them. To brighten things up a bit for Christmas, I used some coloured knobs from RS! (See page 2).

## Next time

The next pedal will be a silicon diode fuzzbox. The PCB is being designed by Mike Grindle to take conventional cheap 16mm pots, so no more nutty struggles. The circuit allows for Silver Tonebender and Big Muff variations to be made. **PE**

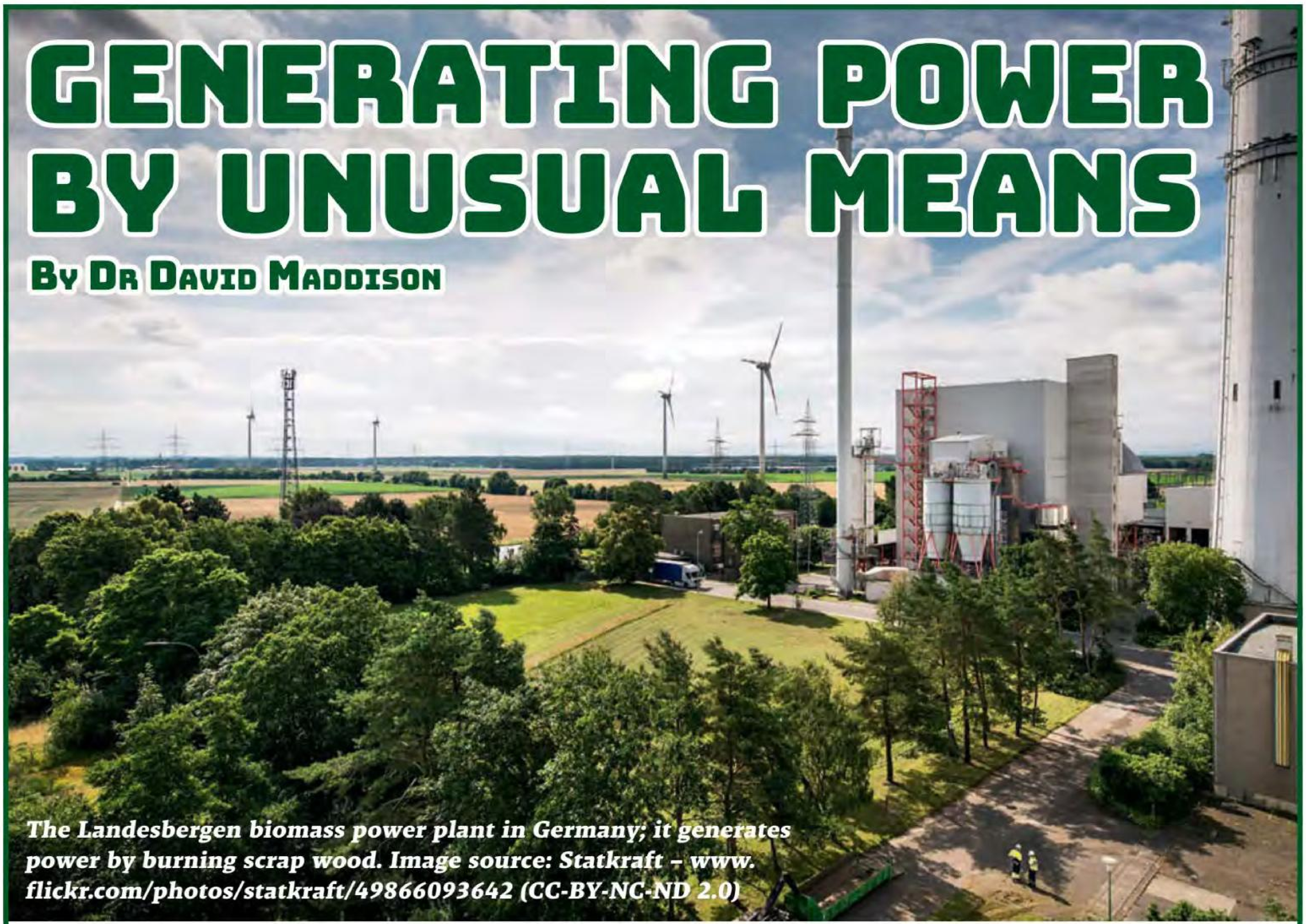


Black
9V
Red
NC switch contact
Centre pin
2.1mm DC connector
Outer
Red
24 SWG link
24 SWG link
0V
S1
Orange
V+
Pre-driver
Input
Output
Yellow
CW
Pink
0V 0V 0V
VR4
47kΩ
Log
Blue
To jack board ground (black)
Ground to switch (green)
Input
Jack sleeve
Ring
Tip
Jack stereo socket
(switch contact not used)
(power switched here)
S2
Grey
Output
Tip
NC
0V
Jack sleeve
S3
Brown
Green
0V
V+
5.6kΩ
0.25W
Effect On
FX output from VR4 wiper
Pink
Grey
Output from jack board
Input from jack board
Violet
S2
Orange
FX input from main board
S1
LED jack board
Brown
S3
Green
0V
Rear of 3PDT switch
(Poles are centre pins
in eacch switch section)

Fig.7: the wiring diagram. This is similar to last month's fuzz box, but uses a conventional negative Earth and with a master volume control added.

# GENERATING POWER BY UNUSUAL MEANS

### By Dr David Maddison



*The Landesbergen biomass power plant in Germany; it generates power by burning scrap wood. Image source: Statkraft – www.flickr.com/photos/statkraft/49866093642 (CC-BY-NC-ND 2.0)*

**Energy is all around us in one form or another, but often in small amounts. Energy harvesting, otherwise known as power harvesting or energy scavenging, is the process of obtaining small amounts of energy from the environment to supply low-power devices.**

**With a few exceptions,** the amounts of power available from energy harvesting are small, and the expense required to obtain that power makes these methods not competitive with grid power, where it is available. However, these tiny amounts of energy can be very useful for powering small devices away from the grid; modern efficient electronics can often run on minimal amounts of power.

This article will cover methods of power generation other than the ones most people are familiar with, like coal and gas generators, nuclear power plants, hydroelectric plants, solar, wind and wave power or burning biomass or waste.

With these alternative power-generating methods, the power available is often on the order of nanowatts to milliwatts. In some cases, it may be possible to generate several watts (or more).

The main applications for energy-harvesting devices include powering IoT devices such as remote sensors, 'wearable electronics', powering biomedical devices (like pacemakers) or charging portable devices like mobile phones.

## Energy harvesting principles

The basic principles and technologies that energy harvesting devices utilise include the following. We will describe their uses when we look at particular implementations.

- Using chemistry, such as in an electrochemical cell.
- Using biochemistry, including the generation of electricity using bacteria or plants.
- Using biomechanical principles, such as utilising bodily movement.
- Using an electret, a dielectric material that maintains electric polarisation after it has been subject to a strong electric field. It is the electrostatic equivalent of a permanent magnet.
- Using electric field gradients, such as causing a fluorescent tube to glow near a power line.
- Using electromagnetic induction to generate electricity by Faraday's Law, the "production of an electromotive force (EMF) across an electrical conductor in a changing magnetic field".
- Capturing electromagnetic radiation from radio waves via an antenna or from light, such as in a solar cell via the photoelectric effect.
- Using electrostatic power generation to produce high voltages at very

low currents. This frequently involves materials rubbing against each other (via the triboelectric effect).

- Using metamaterials, artificial materials with repeating structures that can interact with and manipulate electromagnetic waves in various ways.
- Converting motion to electricity using electromagnetic, electrostatic or piezoelectric effects.
- Using changes in air pressure to expand or contract bellows.
- Using a temperature gradient, such as with a thermoelectric device.
- Using the movement of air, like in a wind turbine.
- Using the movement of water, ie, hydroelectricity.

Below we will cover what energy-harvesting devices and techniques that we have found:

## Watches

Since watches are small, low-powered devices, there has been much interest in energy harvesting to power them.

Self-winding automatic mechanical watches were common before the advent of electronic quartz watches. They had a pendulum activated by swinging one's arm that wound the mainspring. The first credible report of a self-winding pocket watch dates to 1777.

In 1922, the first self-winding wristwatch was invented by John Harwood, and he was awarded Swiss patent 106583 in 1924. The watch was released to the market in 1928 – see Fig.1.

The first solar-powered clock was demonstrated by Patek Philippe at the Basel Fair in 1952! Four hours of light per day was enough to keep the clock running indefinitely. The solar cell drove a motor that wound the mainspring. Patek Philippe went on to make a range of solar clocks; see https://pemag.au/link/abxm

The solar-powered watch was first patented by Timex in 1969, but the first solar watch, the Synchronar 2100, was invented by American Roger Riehl. He partnered with Palo Alto, California based electronics company Ness Time for the project.

The watch (Fig.2) was shown at the RJA Fall trade fair in July 1973 and remained in production until 1983; you can see a TV ad for it at https://youtu.be/mIwxNkGKXb4

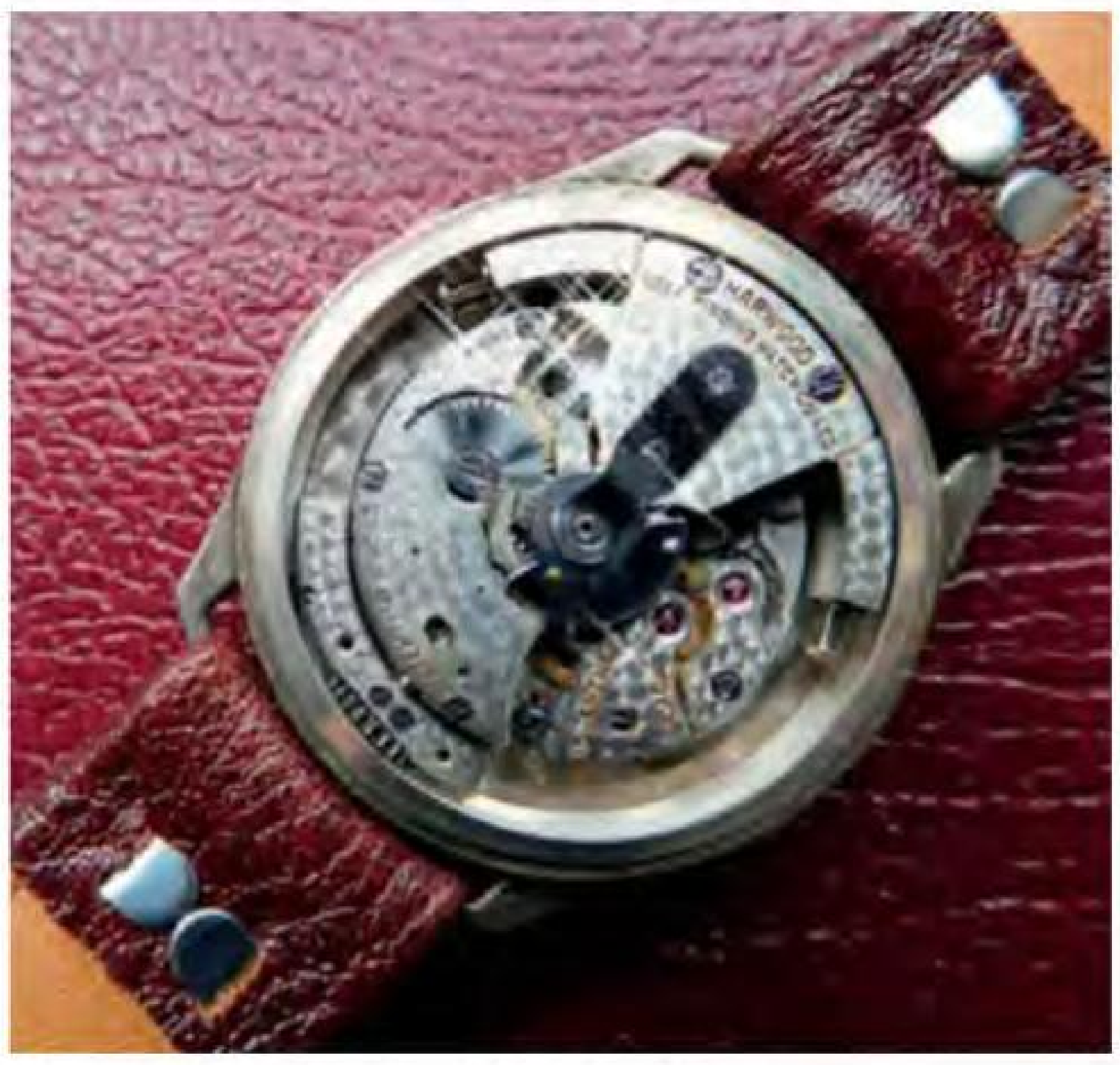


**Fig.1: the first self-winding mechanical watch that harvested energy from the motion of the wearer's arm. *Source: Fratello Watches – https://pemag.au/link/abxy***

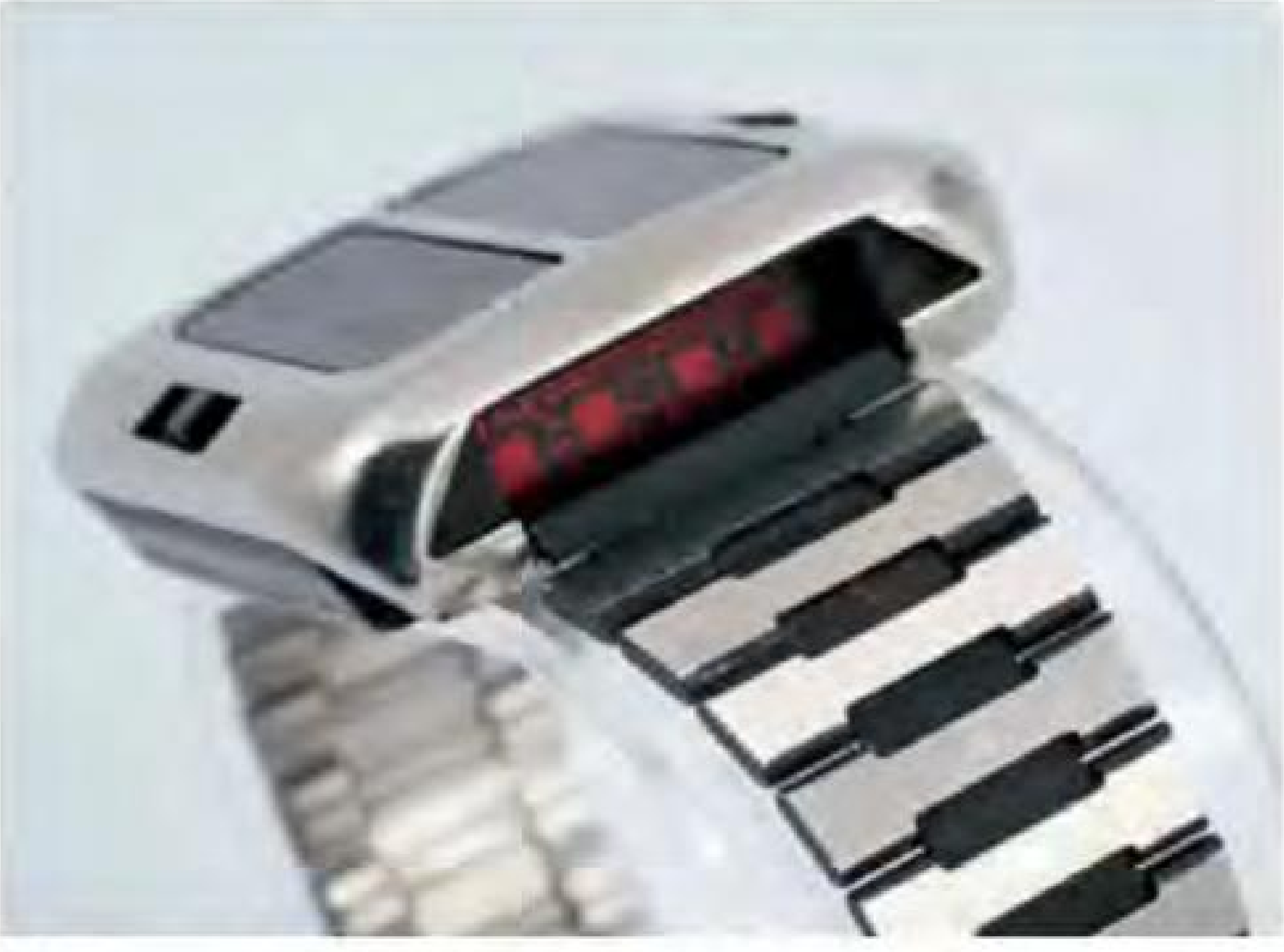


**Fig.2: the Synchronar Sunwatch was the world's first solar-powered watch, released in 1972. *Source: https://solarmuseum.org/cells/synchronar-2100/***

In many modern solar watches, the dial is translucent and the solar cell(s) are hidden beneath it.

Seiko pioneered the so-called automatic quartz watch concept that used a rotating pendulum inside the watch. Instead of winding a spring, it drove a highly-geared miniature generator at up to 100,000 RPM to charge a capacitor or rechargeable battery. Seiko unveiled the technology in 1986 and today sells them under the Kinetic brand.

Seiko still maintains a web page for these watches (https://pemag.au/link/abxn) but we have seen statements that they are being phased out (see https://pemag.au/link/abxo). About eight million have been sold to date. The generator mechanism of the Kinetic watch has been experimentally used to power a cardiac pacemaker in an animal (more on that later).

The PowerWatch uses a Matrix thermoelectric device to power it, in addition to solar energy (www.powerwatch.com). A review of the PowerWatch Series 2 is at https://pemag.au/link/abxp

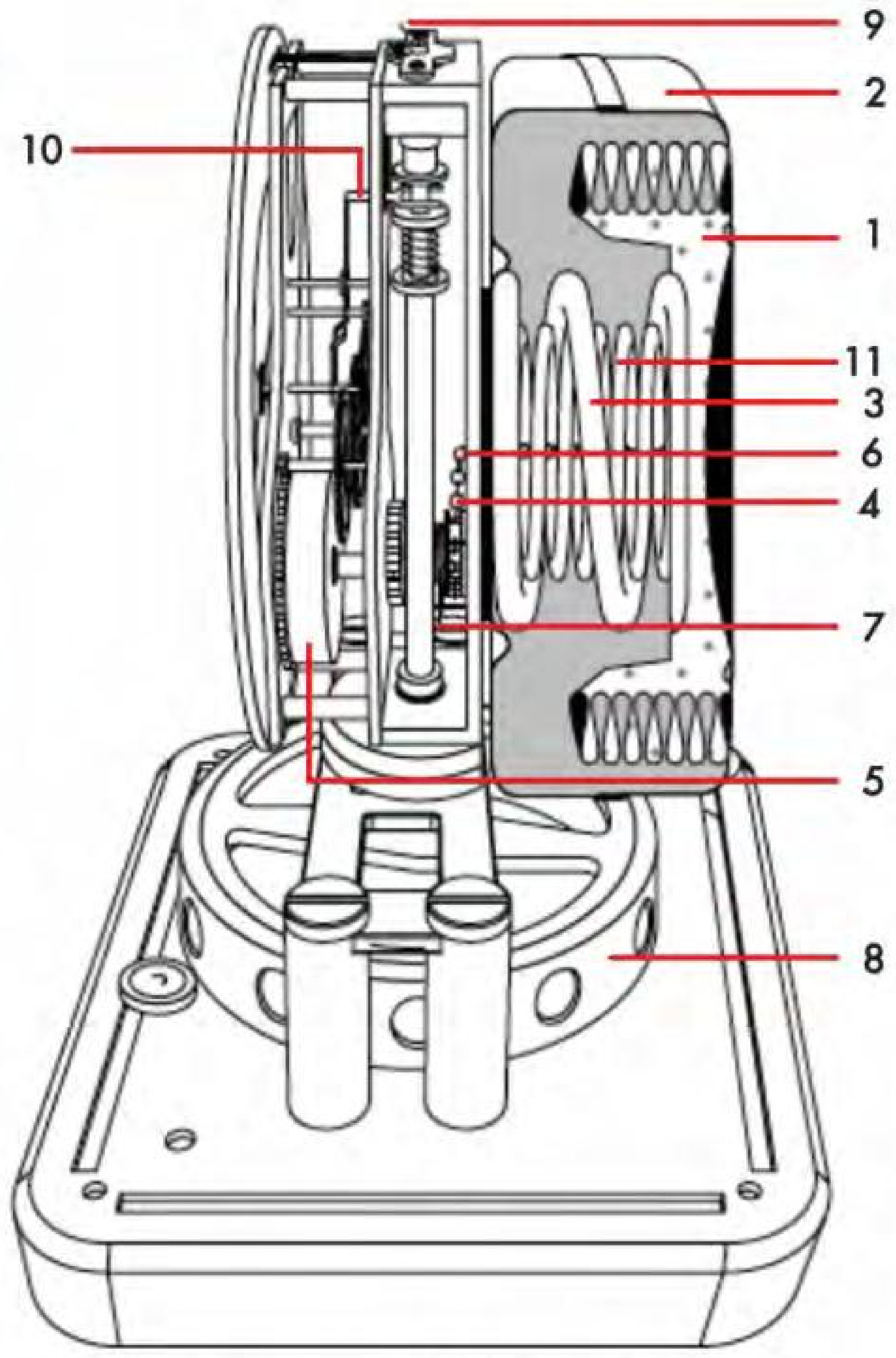


**Fig.3: the Atmos clock mechanism:**

1. **Expansion chamber**
2. **Brass cover**
3. **Balance spring (counterweight)**
4. **Small chain**
5. **Mainspring**
6. **Pulley**
7. **Return spring**
8. **Balance wheel**
9. **Elinvar wire**
10. **Escapement**
11. **Winding spring**

***Original source: Watch Collecting Lifestyle – https://pemag.au/link/abxs***

## Atmospheric & solar clocks

The Atmos is a very expensive clock currently available from Jaeger-LeCoultre that obtains its energy from environmental temperature and pressure changes. Expansion and contraction of liquid and gaseous ethyl chloride in a bellows as the temperature or pressure rises and falls cause a spring to be wound to power the mechanism – see Fig.3.

The Beverly Clock in New Zealand (https://w.wiki/AUgH) has been running since 1864 without winding. However, it did stop a few times, mainly when there was insufficient change in atmospheric pressure or temperature to keep the mechanism wound.

The Long Now Clock (funded by Jeff Bezos; https://longnow.org/clock), being built in the USA, is designed to run for 10,000 years. It uses sunlight falling on a chamber of air to move

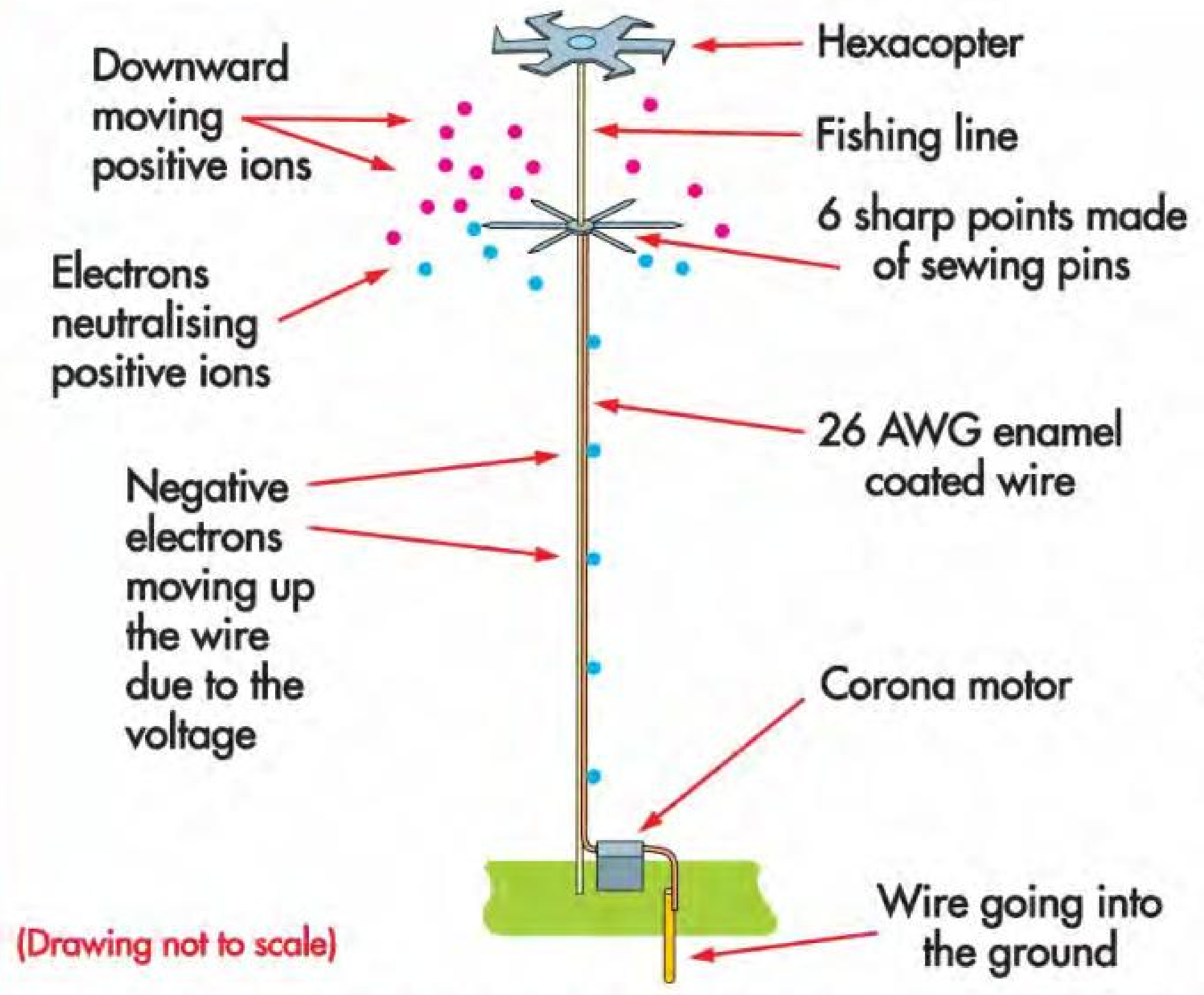

Downward moving positive ions
Electrons neutralising positive ions
Negative electrons moving up the wire due to the voltage
Hexacopter
Fishing line
6 sharp points made of sewing pins
26 AWG enamel coated wire
Corona motor
Wire going into the ground
(Drawing not to scale)

**Fig.4: harvesting atmospheric electricity to run an electrostatic motor. This type is called a corona motor.** ***Original source: Rimstar –*** [***https://pemag.au/link/abx7***](https://pemag.au/link/abx7)

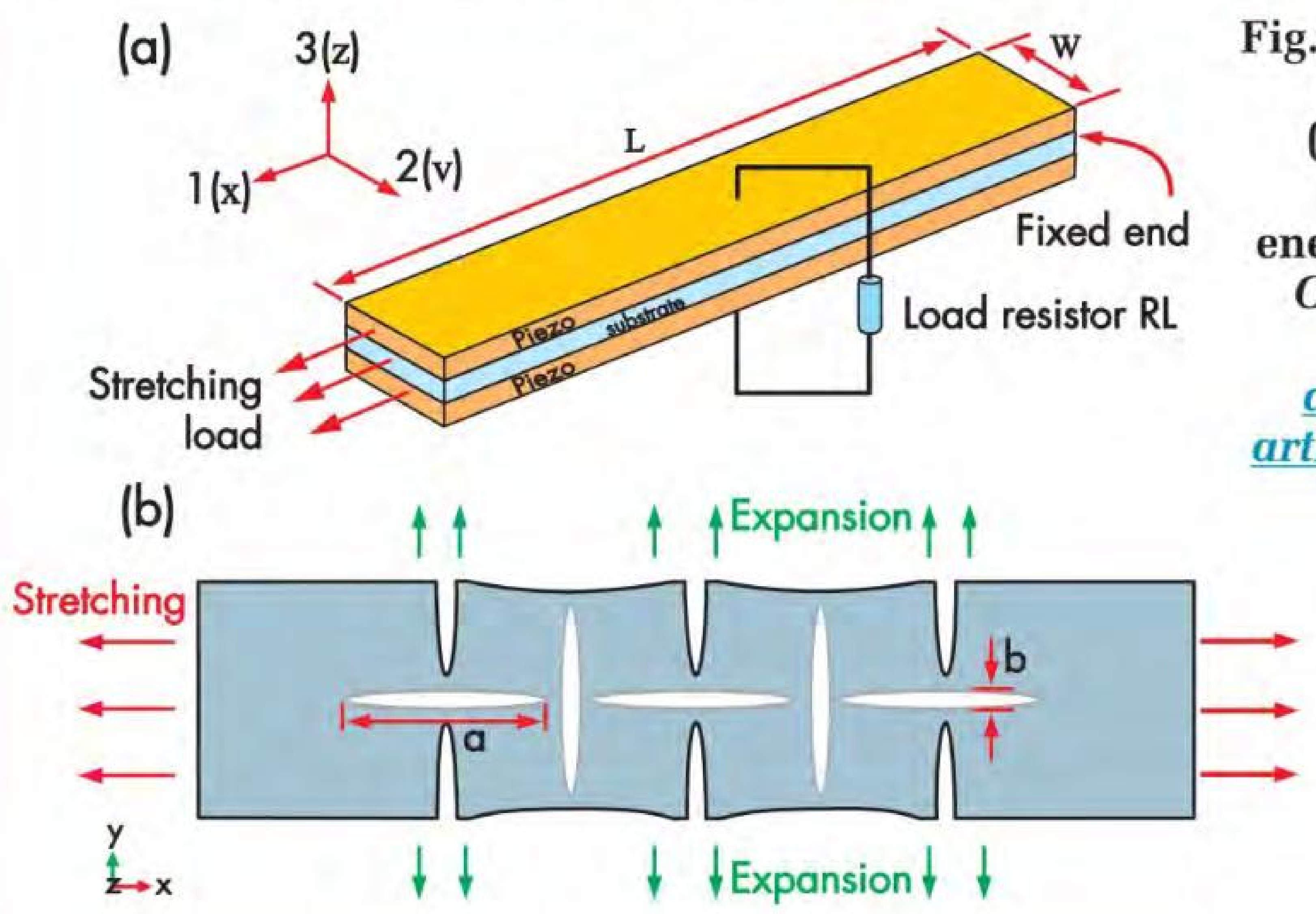


**Fig.5: conventional (a) and auxetic (b) piezoelectric bimorphs for energy harvesting.** ***Original source:*** [***https://pubs.aip.org/aip/adv/article/7/1/015104/240312/***](https://pubs.aip.org/aip/adv/article/7/1/015104/240312/)

a cylinder, which provides enough winding force to keep the pendulum going. It is also used to synchronise the clock to solar noon. So, in a sense, it is solar powered, although it does not use a photovoltaic panel.

## Atmospheric electricity

There is a substantial electric field gradient in the atmosphere, so an electrostatic motor can be made to turn by having one electrode high in the air with the other at a lower level (see Fig.4). The power is meagre; at most a current of a few microamps can be drawn.

For more on this, see the panel in this article on Hermann Plauson (page 66), the video titled "How Powering with Atmospheric Electricity Works" at **https://youtu.be/2rVdEhyMR6A** and the web page at **https://pemag.au/link/abx7**

## Piezoelectric energy

Piezoelectricity involves the production of electrical energy from mechanical strain. Examples of sources of strain include motion, sound and vibration. The power generated is typically minimal, milliwatts or less. Piezoelectric materials include ceramics like quartz crystals and, more recently, piezoelectric polymers like polyvinylidene fluoride (PVDF) – see Fig.6.

An example of a piezoelectric energy harvester is shown in Fig.7.

Some piezoelectric substances are also pyroelectric. These crystals are naturally electrically polarised and produce a voltage when heated or cooled. This could be used for energy harvesting over a day by taking advantage of the natural changes in ambient temperature.

Auxetic materials are artificially-structured metamaterials that expand in width rather than contract when stretched. Conversely, when subject to compression, they reduce in width. It has been proposed that auxetic materials could increase the energy-harvesting efficiency of piezoelectric devices, as shown in Fig.5.

In that figure, (a) shows a conventional piezoelectric bimorph, which can generate power mainly in the stretching direction, while (b) represents a bimorph of auxetic construction.

This can generate power simultaneously in both the stretching and transverse directions, resulting in an expected power increase of 176%.

That is because it has increased power output in the transverse direction, as it can generate more stress in that direction, and the power output is proportional to the applied stress.

Clothing has been proposed that incorporates piezoelectric materials to generate power for powering or charging devices. Such fabric utilises nanofibres and is said to be stretchable and breathable. See **https://pemag.au/link/abxe**

## Thermoelectricity

Thermoelectricity involves the production of an electric current due to a thermal gradient between two dissimilar electrical conductors. A typical example of a device that utilises this effect is a thermocouple, although it produces tiny amounts of power at very low voltages.

Peltier devices (Fig.8) also utilise this effect but with many more thermoelectric junctions. When a current is applied, it can move heat towards or away from an object. Alternatively, when a temperature differential is applied, it can generate a voltage and current, and thus be used for energy harvesting.

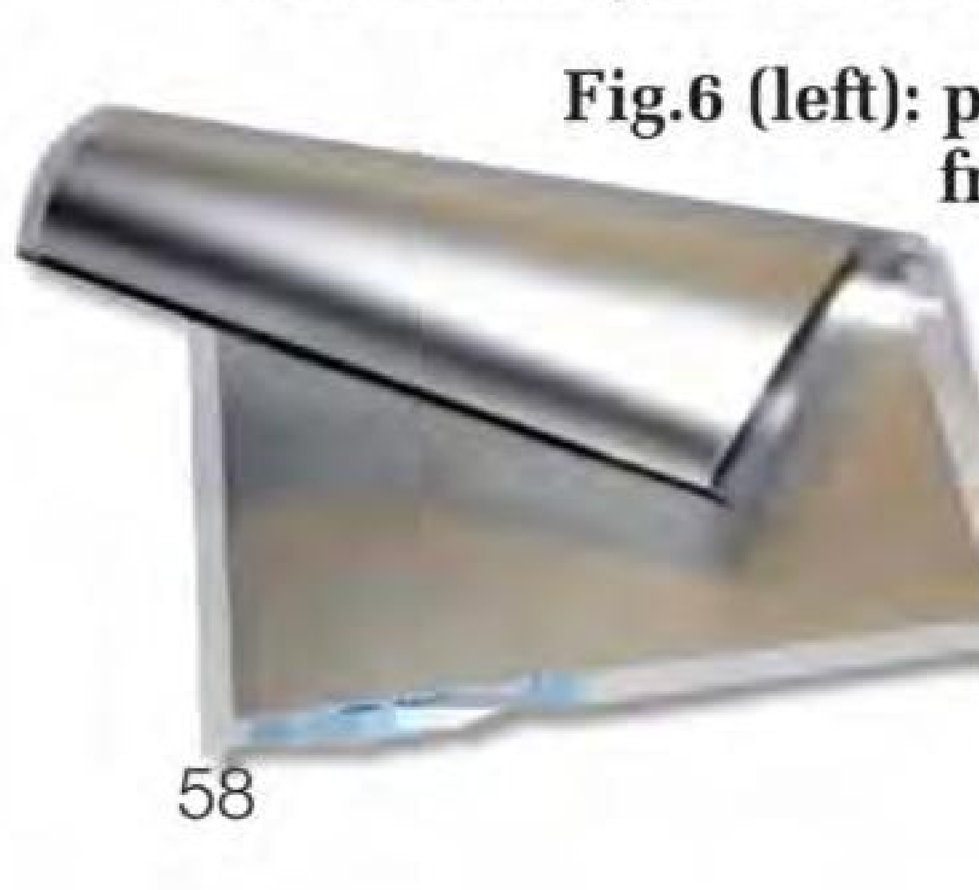


**Fig.6 (left): polyvinylidene fluoride (PVDF), a piezoelectric material, with deposited electrodes from a commercial supplier.** ***Source:*** [***www.he-shuai.com/pvdf-piezo-film***](https://www.he-shuai.com/pvdf-piezo-film)

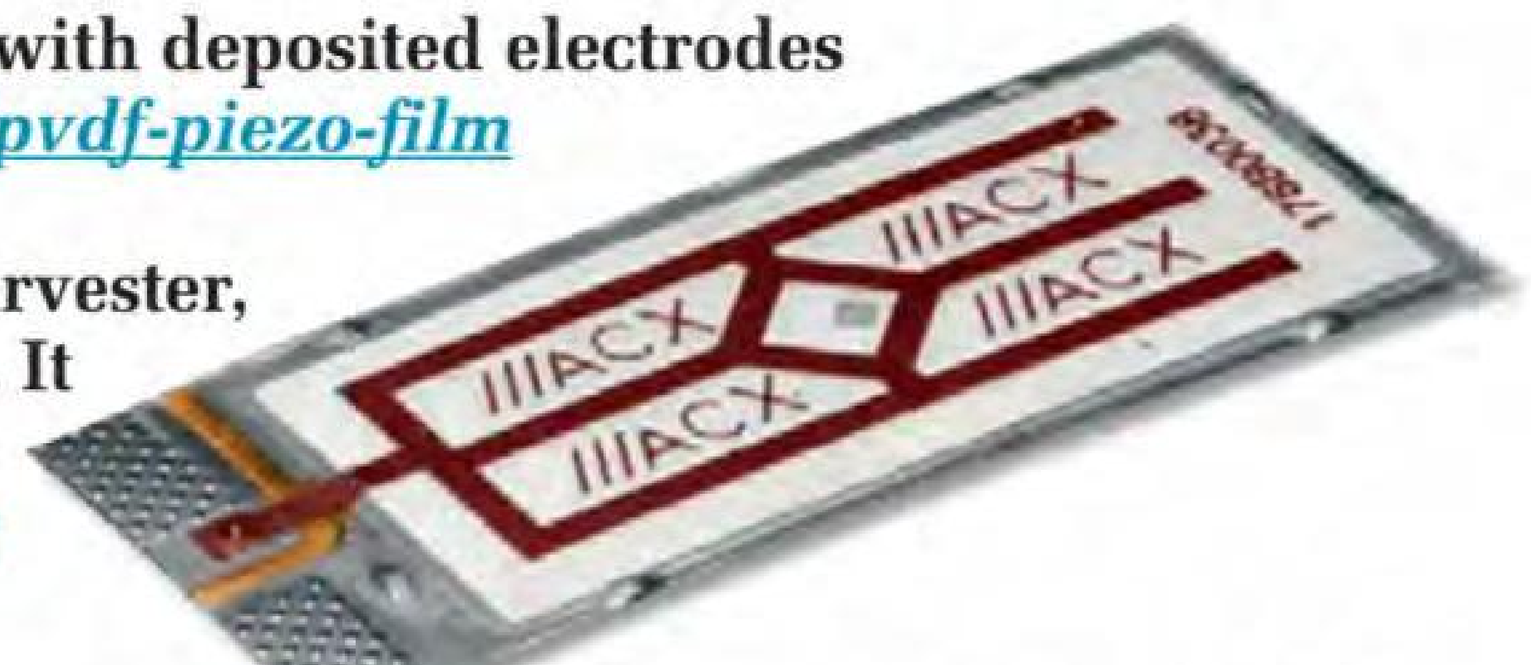


**Fig.7 (right): a commercial piezoelectric energy harvester, model S118-J1SS-1808YB (from https://piezo.com). It can produce up to 0.7mW.** ***Source: Piezo S118-J1SS-1808YB –*** [***https://pemag.au/link/abxv***](https://pemag.au/link/abxv)

# Unconventional Energy Harvesting

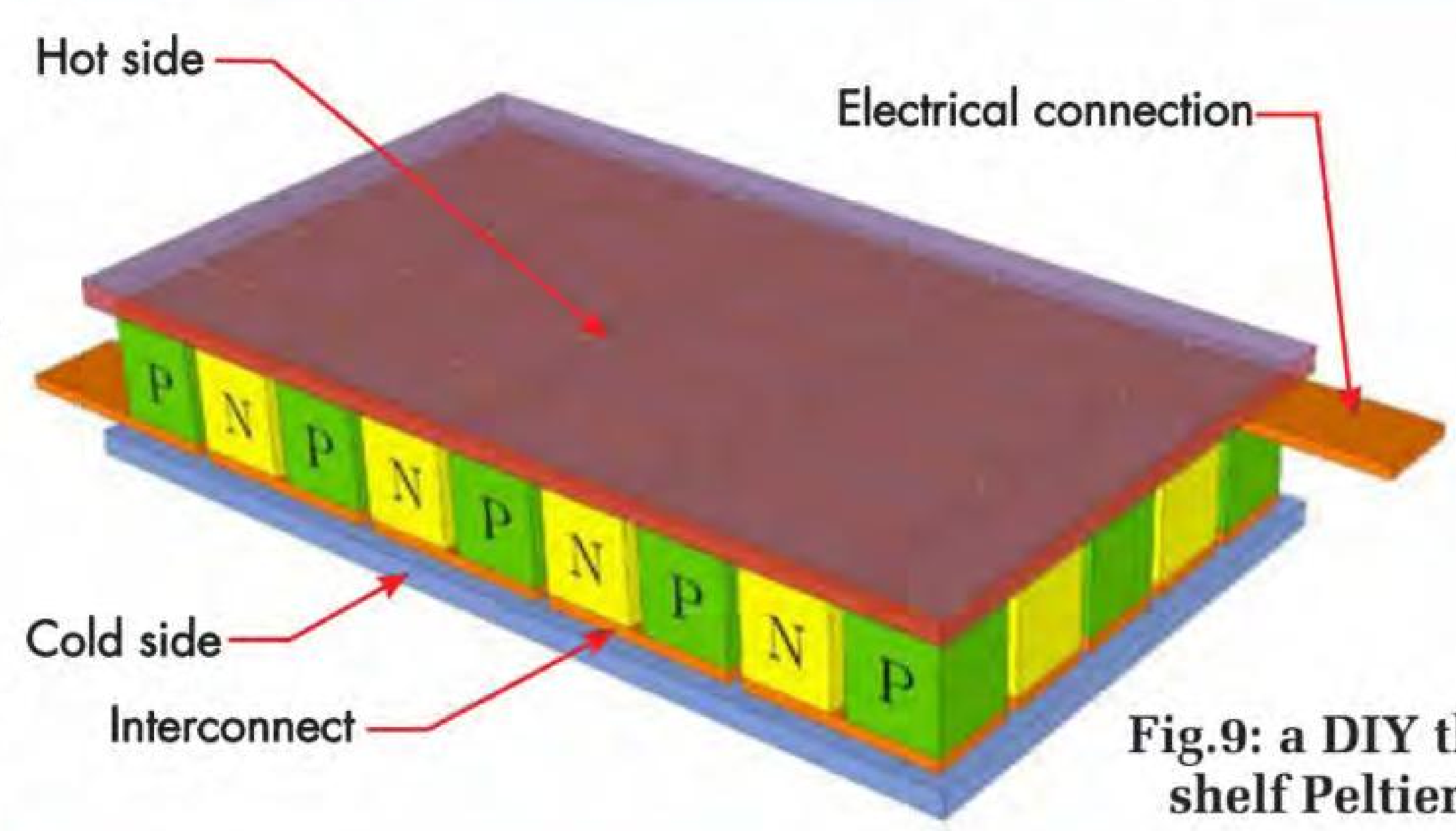


**Fig.8: a Peltier device. It uses a combination of p-type and n-type semiconductor materials to create thermoelectric junctions. They are connected electrically in series and thermally in parallel.** ***Original source:*** ***[https://w.wiki/AUjV](https://w.wiki/AUjV)***

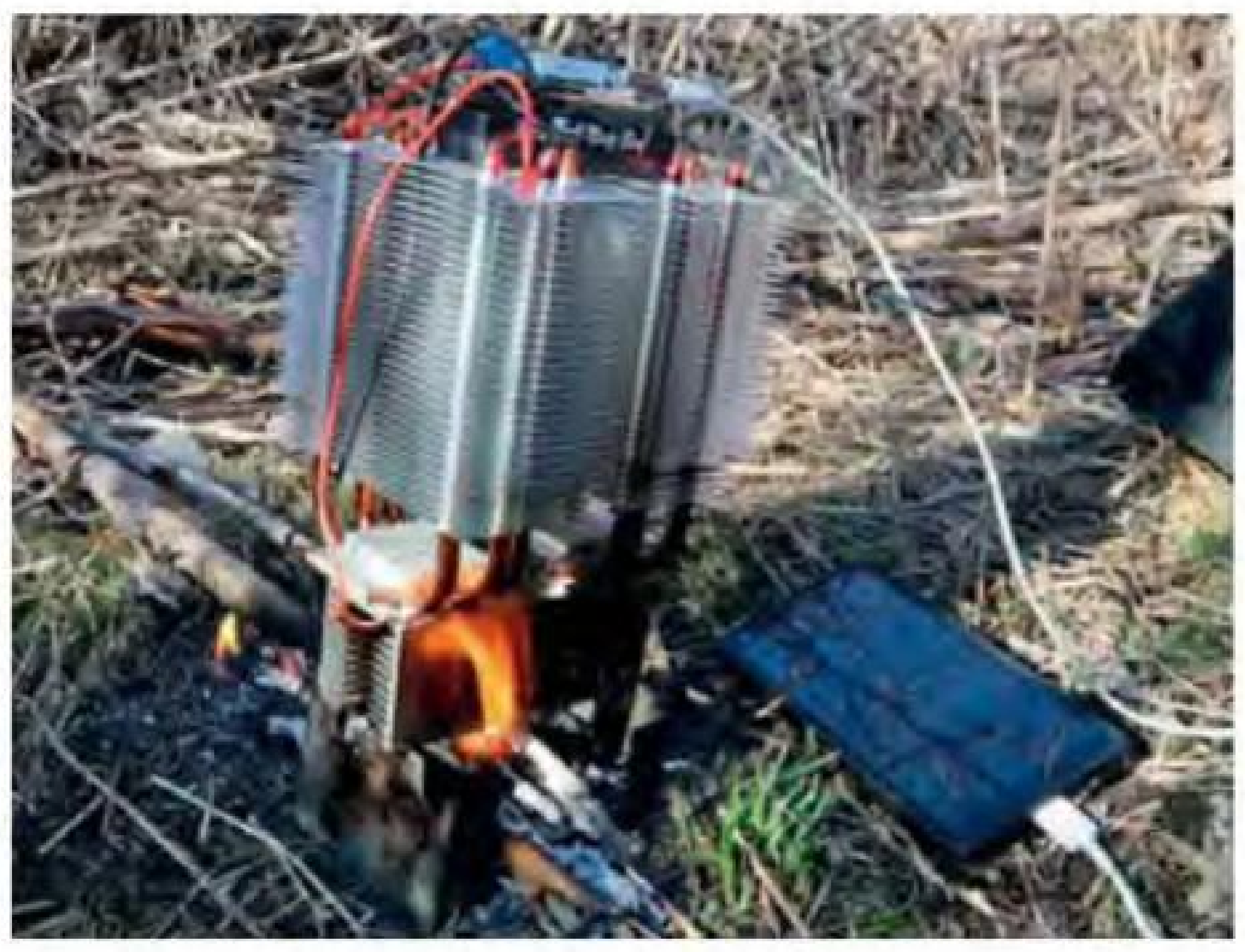


**Fig.9: a DIY thermoelectric generator using an off-the-shelf Peltier device, heatsink and other components.** ***Source:*** ***[https://youtu.be/x9a2rB-xWkY](https://youtu.be/x9a2rB-xWkY)***

Electricity can be generated from a campfire using thermoelectric principles. Fig.9 shows a Peltier device attached to a heatsink that can generate power from a fire. The CampStove 2 from BioLite can produce up to 3W to power or charge USB devices (see **https://pemag.au/link/abxa**).

The MATRIX Prometheus Thermal Energy Harvesting Module produces power by exploiting small environmental temperature differences, using the thermoelectric effect. The most powerful Prometheus device, the PRMT02-34465, produces up to 14mA (**www.matrixindustries.com/0234465**).

This technology is used to power the MATRIX Perceptive Health Monitor, their Proximity Sensor and the PowerWatch (**www.powerwatch.com**).

## Stirling engines

A Stirling engine is a type of heat engine that can function with very small heat differences and thus can be used for energy harvesting from low-grade heat sources – see Fig.10.

The Stirling engine can be connected to a generator to produce electricity. Stirling engines have been proposed by NASA to produce power on a future mission to Mars (see also the 2020 Mission to Mars article in the July 2021 issue of Silicon Chip magazine).

## Energy from bacteria

Some exotic bacteria exchange electrons with the environment ('extracellular electron transfer' [EET]), so theoretically, they could be used to produce electricity. Mechanisms from these exotic bacteria have been genetically engineered into common E. coli bacteria. Such an approach could be used to convert wastewater effluent streams into electricity.

However, this is very early work and practical applications are a long way off. The work was published at **https://pemag.au/link/abxd**

Also see the video titled "Scientist engineered bacteria to generate electricity from wastewater" at **https://youtu.be/beL_qlsmNQ8**

## Power from plants

A common experiment for children is (or used to be) to use a lemon, potato or other fruit or vegetable to make a basic electrochemical cell (see Fig.11). Pieces of different metals, such as zinc and copper, are used as electrodes, while the juice of the fruit or vegetable acts as the electrolyte. One such cell might produce 0.9V at 1mA. Several lemons can be connected in series to power one LED.

A fun experiment was once performed to see if a 1000-lemon battery could start a car. See the video titled "Can a battery made from 1000 lemons start a car?" at **https://youtu.be/4f2wsQkQ71o**

Light can be turned into electrical energy via the photosynthesis mechanism using bio-photoelectrochemical cells (BPECs). This early work is described in the scientific publication at **https://pemag.au/link/abxl**

## Biomechanical energy from the human body

Raziel Riemer and Amir Shapiro calculated the energy available from the

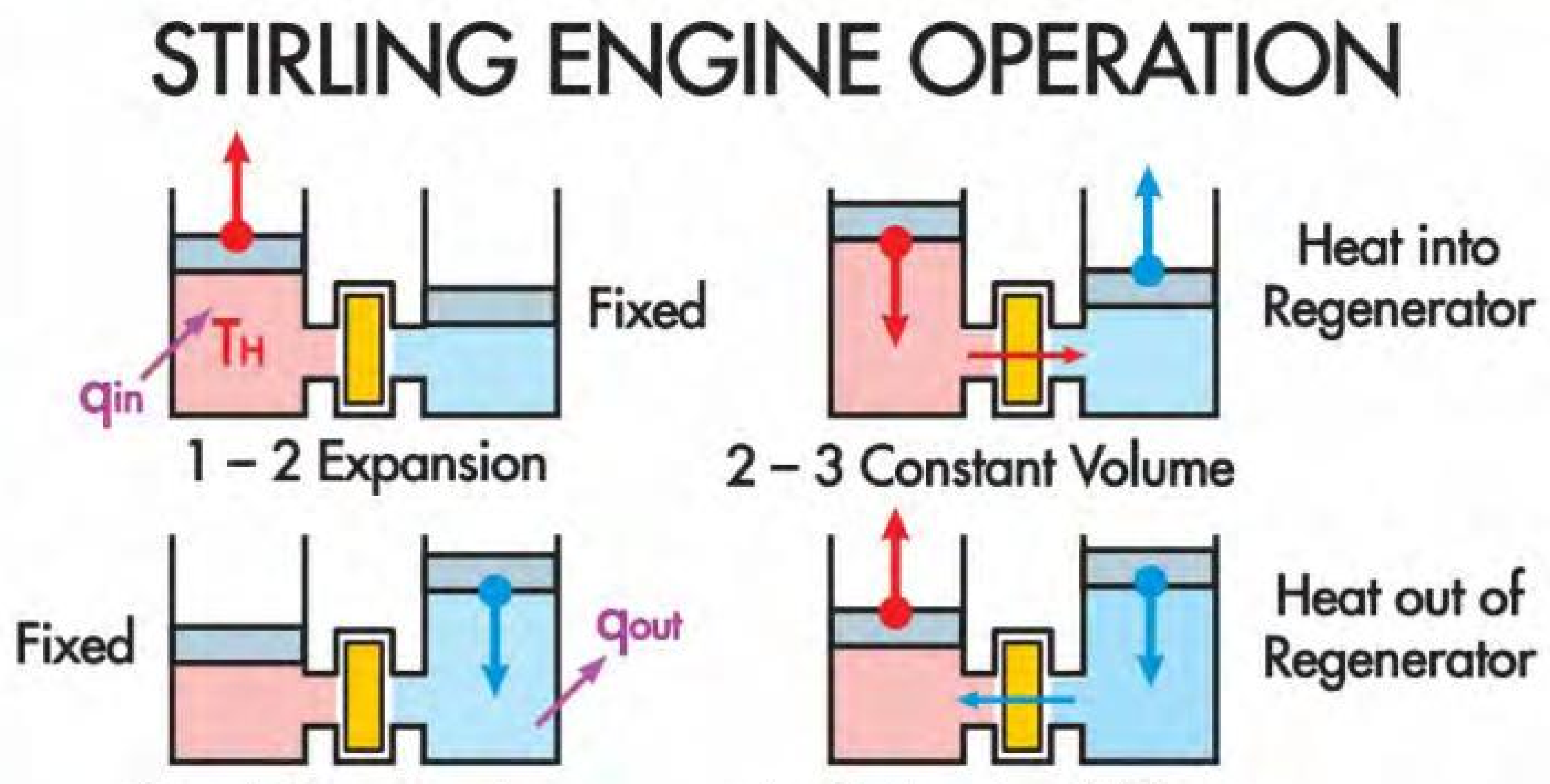


**Fig.10: the operating cycle of a Stirling engine, which can run from relatively low temperature differentials and could be used as part of a generator.** ***Original source:*** ***[https://youtu.be/hbfkbcdw_OM](https://youtu.be/hbfkbcdw_OM)***

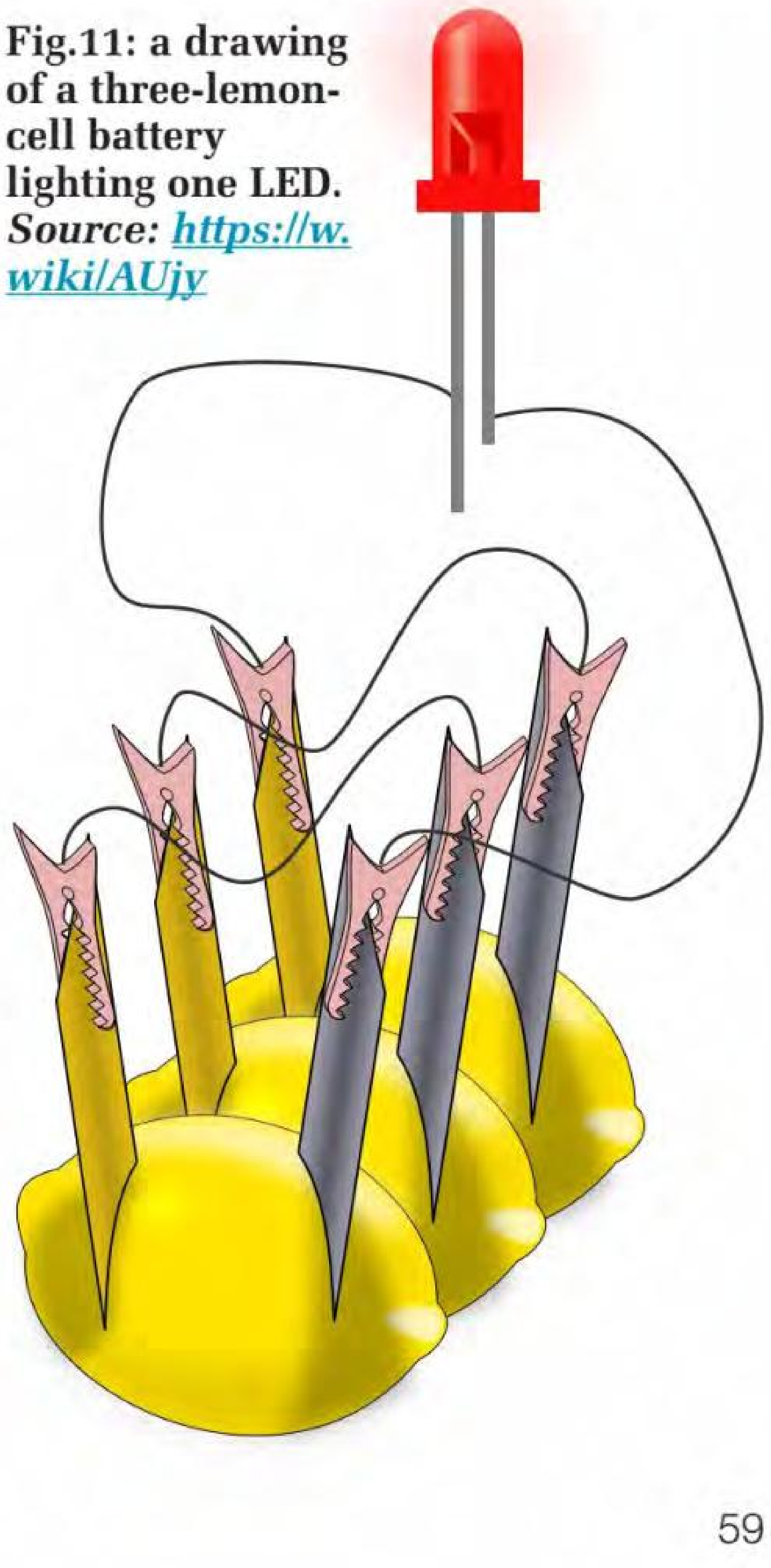


**Fig.11: a drawing of a three-lemon-cell battery lighting one LED.** ***Source:*** ***[https://w.wiki/AUjy](https://w.wiki/AUjy)***

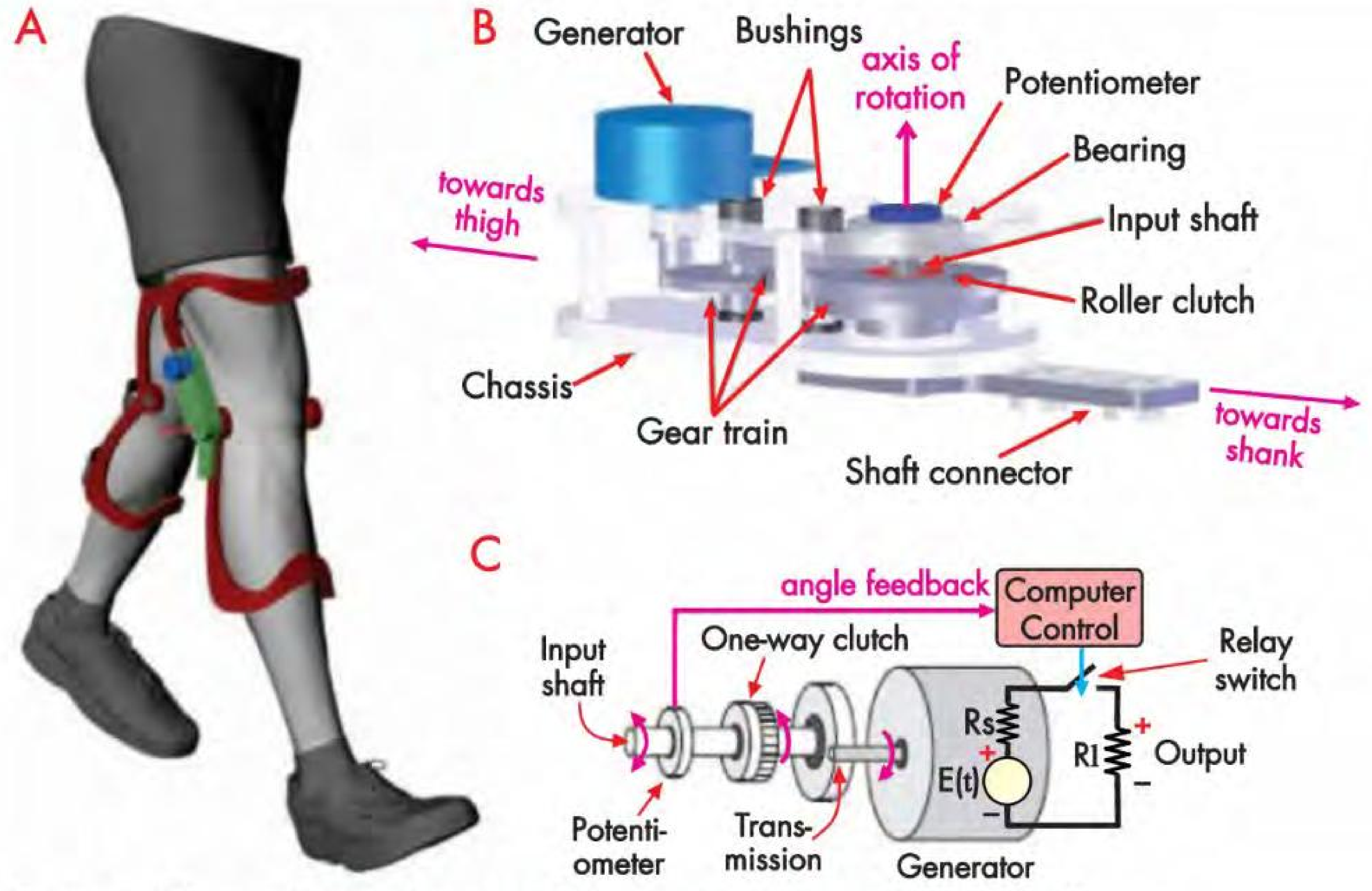
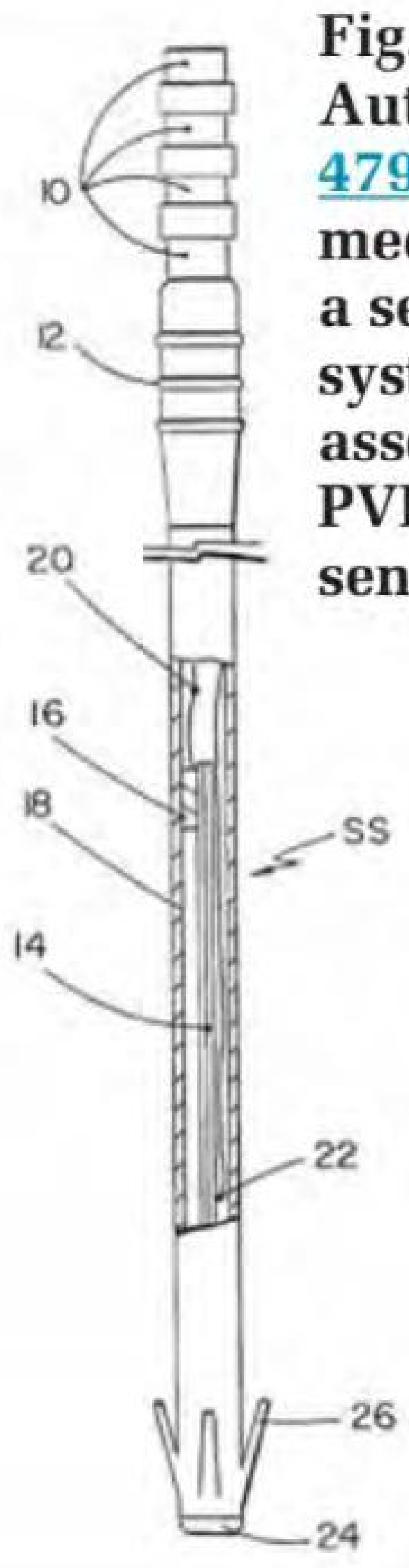

A
B
Generator
Bushings
axis of rotation
Potentiometer
Bearing
towards thigh
Input shaft
Roller clutch
Chassis
Gear train
Shaft connector
towards shank
C
angle feedback
Computer Control
Input shaft
One-way clutch
Relay switch
Rs
E(t)
Rl
+
−
Output
Potenti-ometer
Trans-mission
Generator

**Fig.12: a biomechanical energy-harvester that mounts on the knee. *Original source: www.researchgate.net/publication/51078340***


10
12
20
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18
SS
14
22
26
24

**Fig.13: an image from the Author's 1989 US Patent 4798206 for "Implanted medical system including a self-powered sensing system" showing an assembly of piezoelectric PVDF polymer as the sensing element (#14).**

motion of an 80kg human body under various circumstances (https://pemag.au/link/abx9) and found the following power available:

- heel strike: 2-20W
- ankle motion: 67W
- knee motion: 36W (see Fig.12)
- hip motion: 38W
- movement of centre of mass: 20W
- elbow motion: 2W
- shoulder motion: 2W

They point out that the typical human body consumes the equivalent of 800 AA cells (which would weigh 20kg) by burning just 200g of fat.

## Cardiac pacemakers

A rough estimate for the energy consumption of an implantable cardiac pacemaker is around 10-100µW. Over 5-10 years, that amounts to about 0.5-2Ah. The low power level makes it an ideal target for energy harvesting. That would mean, instead of the pacemaker having to be replaced when the battery goes flat, it could be powered indefinitely.

Fig.13 shows one of the Author's US Patents from 1989 for a pacemaker "self-powered sensing system". It generates electrical signals from the heart's motion using a polyvinylidene fluoride (PVDF) piezoelectric film.

A Seiko Kinetic watch mechanism was also demonstrated experimentally to generate power for a pacemaker; see https://pemag.au/link/abxt

Another option for powering a pacemaker is an 'inertia-driven triboelectric nanogenerator' (I-TENG), as described at https://pemag.au/link/abxb

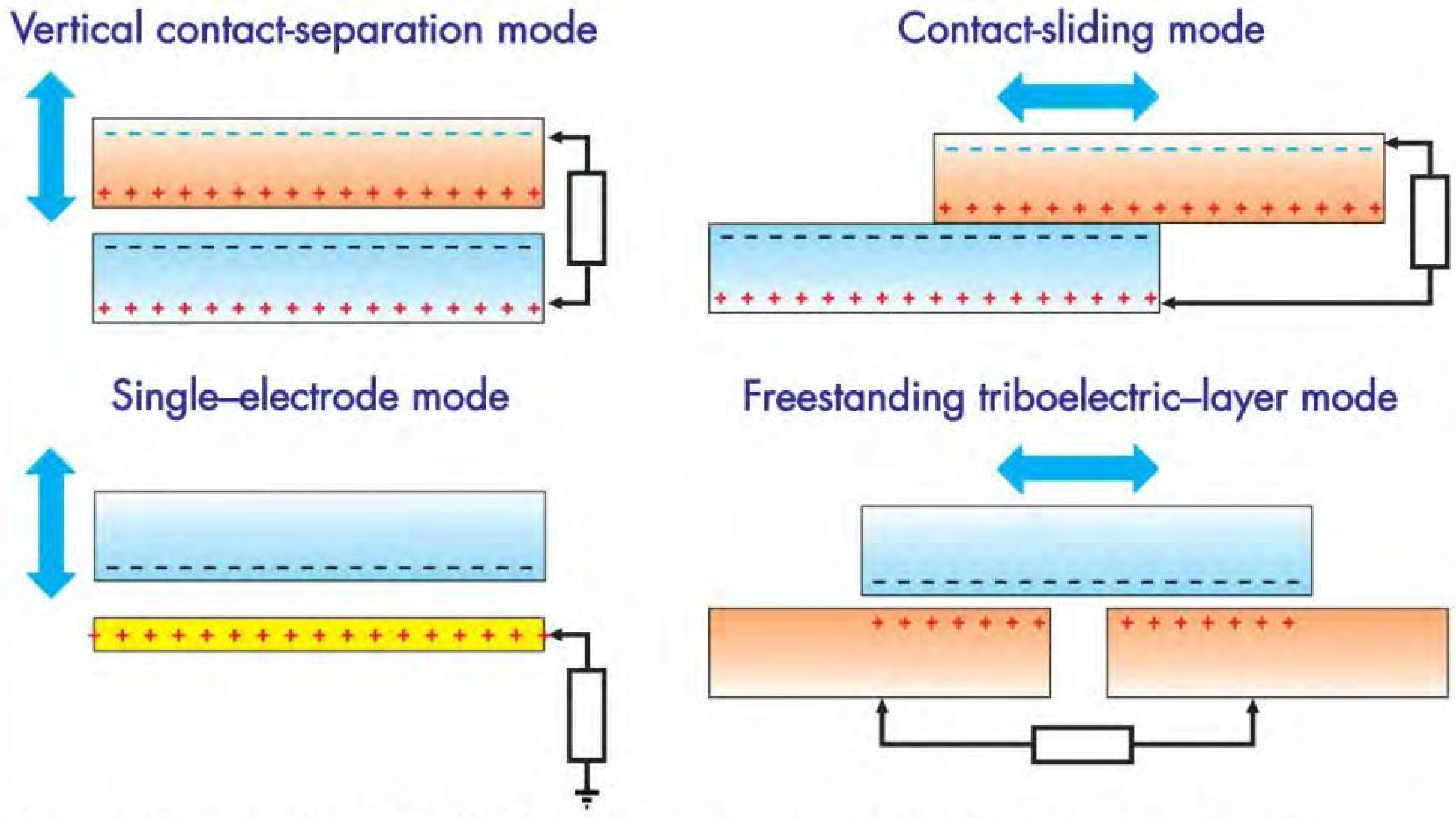

Vertical contact-separation mode
Contact-sliding mode
Single–electrode mode
Freestanding triboelectric–layer mode

**Fig.16: four modes of triboelectric generators. *Original source: www.researchgate.net/publication/322251641***

## Triboelectricity

The triboelectric effect is electric charge transfer due to two objects rubbing together. For example, a shoe rubbing on a carpet can result in a static electricity shock to the wearer when they touch a grounded object.

A 'drinking bird' toy can be turned into a 'triboelectric hydrovoltaic generator' using two effects. A temperature differential powers the bird, while triboelectricity is used to generate power.

Experiments demonstrated such a generator powering items like liquid crystal displays, temperature sensors and calculators. For further details, see https://pemag.au/link/abxc

A triboelectric nanogenerator (TENG) is an energy-harvesting device that generates an electric charge using the triboelectric effect involving a periodic contact or sliding motion – see Fig.16. Low currents are produced at high voltages.

## Electret power generators

An electret is the electrostatic equivalent of a permanent magnet, and a moving electret can be used to produce power similarly to a magnet. You would probably be familiar with electrets in electret microphones; they serve to bias on the FET within the microphone capsule in the absence of an external voltage source.

An electret-based power generator has been demonstrated using microelectromechanical (MEMS) principles

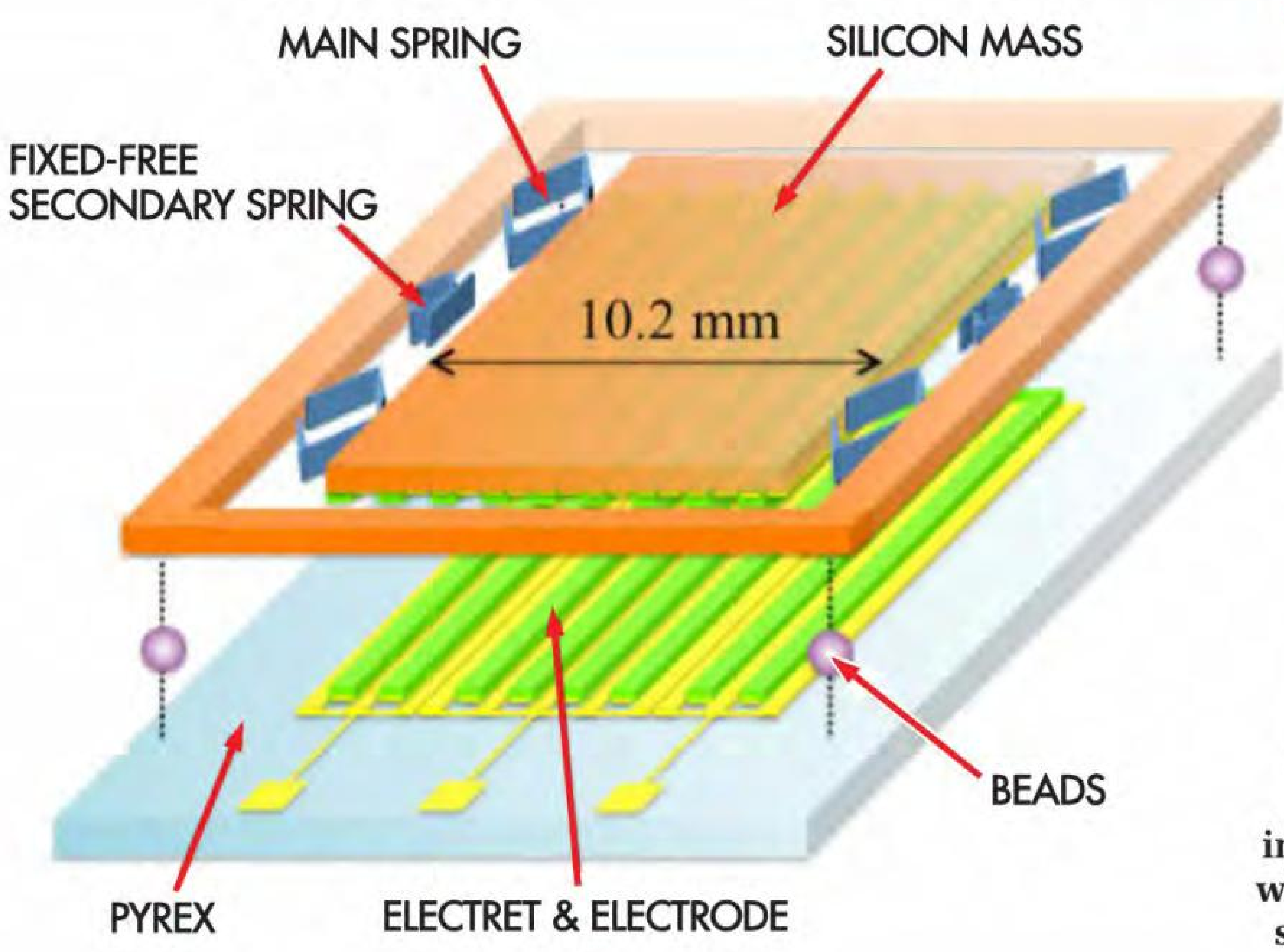


**Fig.14: an energy-harvesting prototype that converts vibration into electricity using MEMS technology and the electret principle. *Original source: [www.mesl.t.u-tokyo.ac.jp/en/research/electret.html](www.mesl.t.u-tokyo.ac.jp/en/research/electret.html)***

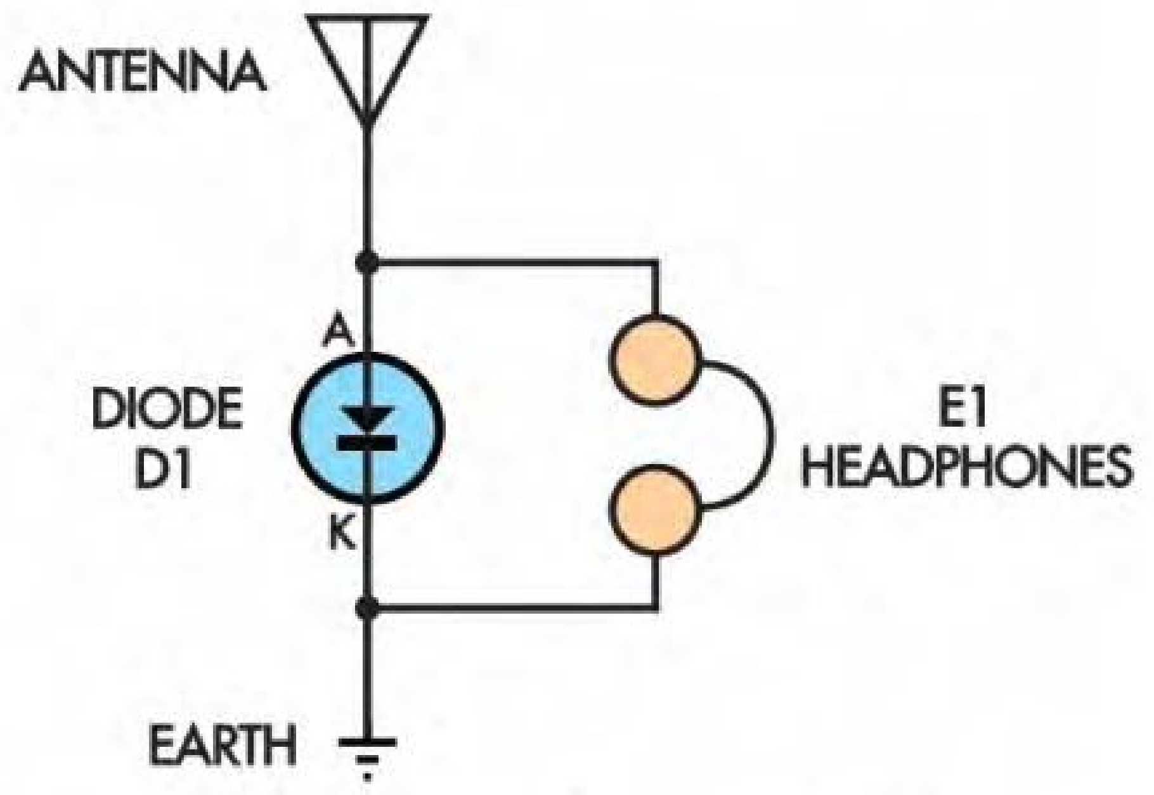


**Fig.15: the circuit of the simplest possible crystal radio using a diode, long wire antenna and high-impedance headphones. Lacking a tuned circuit, it will receive all stations at once, but in practice, the strongest station will probably drown out the rest. *Original source: [https://w.wiki/AUjt](https://w.wiki/AUjt)***

as described at [https://pemag.au/link/abxi](https://pemag.au/link/abxi) (see Fig.14). The prototype produced 6μW from an acceleration of 13.73m/s² at 40Hz

## Power from radio waves

Crystal radios were made by children back in the day and could obtain useful radio reception without a battery. They were powered by harvesting the energy of the radio wave itself – see Fig.15.

RF energy can also be harvested for other purposes using a tuned antenna and a rectifier that works at the desired frequency. They must be close to a source of RF, such as a WiFi router.

Commercial modules to harvest RF energy include the Powercast P2110B, which converts RF to DC. It is optimised to absorb energy in the 850-950MHz range and can provide a regulated output of up to 5.5V – see Fig.17.

Some YouTube videos demonstrate harvesting small amounts of power from commercial radio stations. The author of the following video manages to light ten LEDs, although he is only 1.6km from the radio station: "Free Energy From Radio Waves ([https://youtu.be/_pm2tLN6KOQ](https://youtu.be/_pm2tLN6KOQ)). Fig.18 shows another RF-energy-harvesting circuit.

Peter Parker looks at whether you can harvest enough power to drive a speaker with a crystal set next to a commercial radio station transmitter in the video titled "Crystal set under a 100kW radio station: How does it sound?" at [https://youtu.be/xglEsaNkPSA](https://youtu.be/xglEsaNkPSA)

## Würth Elektronik's energy harvesting evaluation kit

Würth Elektronik ([www.we-online.com/en](www.we-online.com/en)) offers an energy-harvesting evaluation kit with several energy-harvesting options – see [https://pemag.au/link/abxq](https://pemag.au/link/abxq)

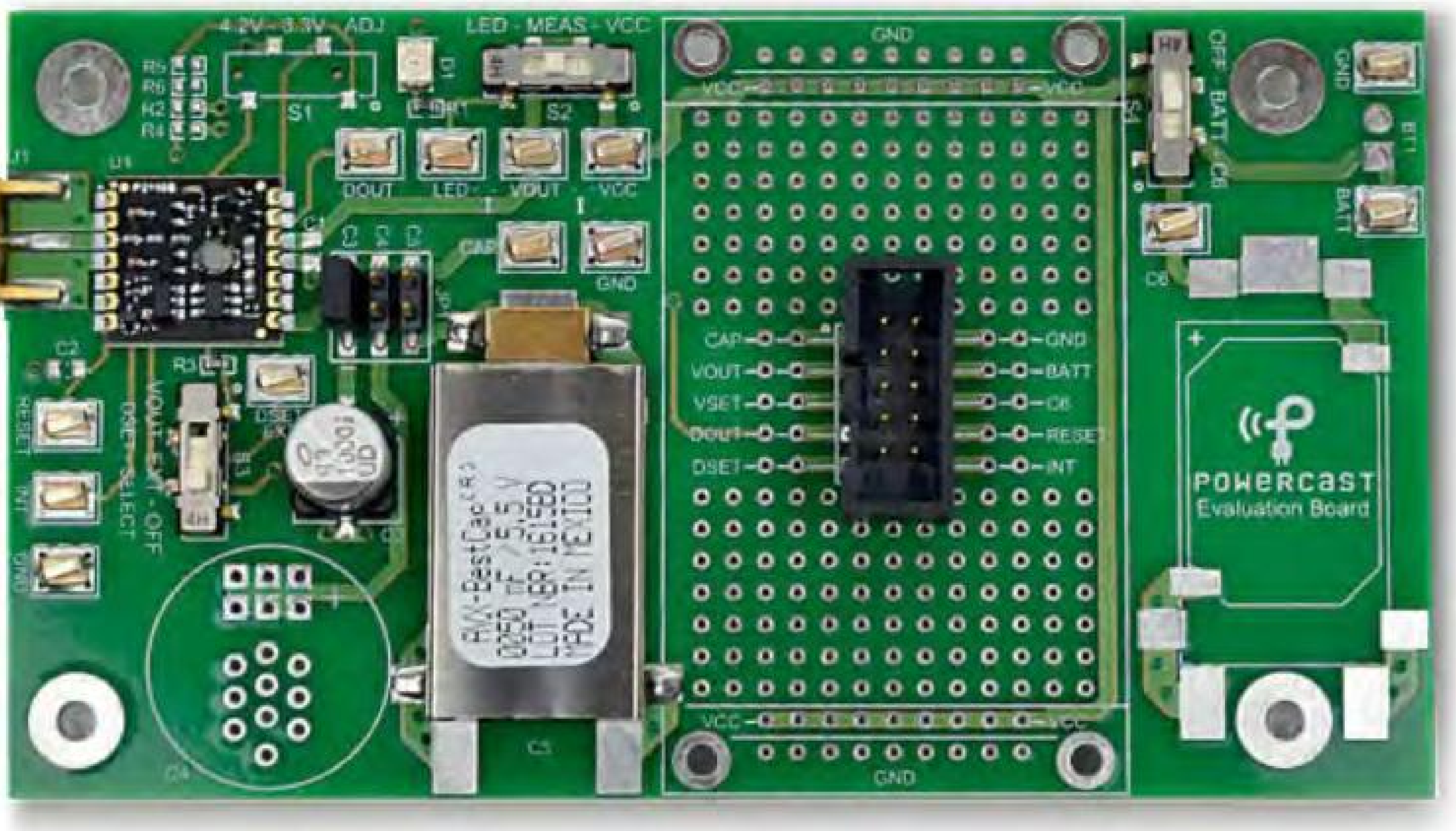
## Earth batteries

An Earth battery (or, more correctly, cell) is made by inserting two dissimilar metal electrodes in the ground. Zinc and copper are two metals that can be used as electrodes. The soil must be moist for the cell to work. Multiple cells can be connected to make a battery.

It is not "free" energy because, as with any cell, one or both of the electrodes will eventually be consumed or deteriorate. Also, the ions in the soil will eventually be depleted, and a new location will have to be selected.

The first Earth battery was invented

**Fig.17: a P2110B energy harvester module on a Powercast evaluation board. The module needs a suitable antenna and capacitor to operate. *Source: All About Circuits – [https://pemag.au/link/abxw](https://pemag.au/link/abxw)***



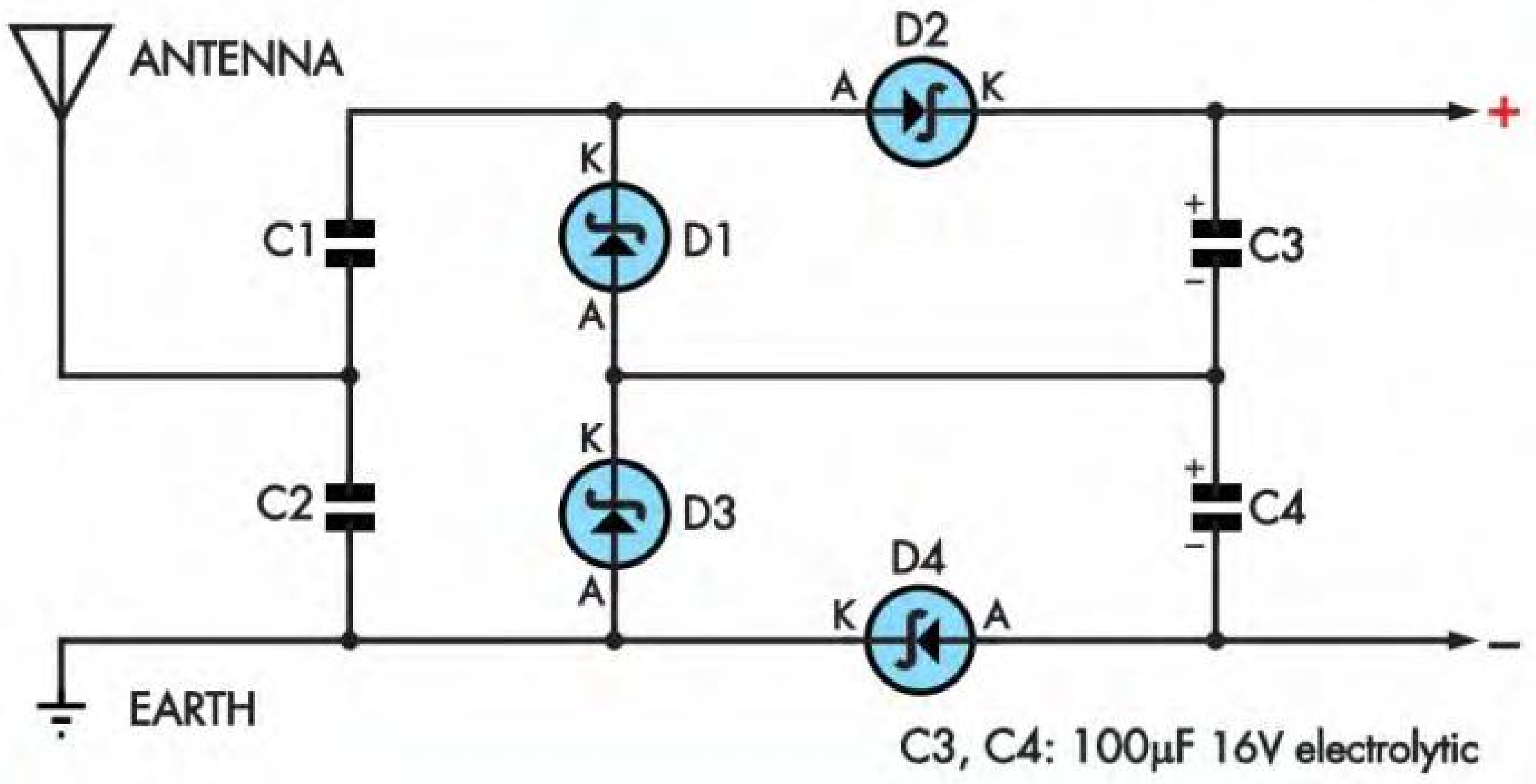


**Fig.18: an energy-harvesting circuit for ambient radio waves, although the amount of energy collected is tiny. *Original source: [https://youtu.be/XpLCK88nVgU](https://youtu.be/XpLCK88nVgU)***

by Alexander Bain in 1841; he used zinc and copper electrodes.

From an electrochemical point of view, there is nothing unusual about an Earth battery, apart from the medium being the ground rather than a more conventional container such as a battery case.

Power harnessed from Earth batteries should not be confused with telluric currents. Still, telluric currents might contribute to the overall EMF of the cell if the electrodes are sufficiently far apart.

## Telluric currents

Telluric currents are electrical currents within the Earth or sea induced by magnetic disturbances from various sources, both natural and artificial. That includes space weather, such as the solar wind, sunspots and their interaction with the ionosphere.

They can be a problem for underground and undersea cables and buried pipelines. As they can be influenced by the sun, they vary during the daily solar cycle.

In the 1800s, problems in telegraph operation were recognised to be related to telluric currents due to sunspots. In 1903, W. Finn reported in Scientific American that an EMF of 768V with a current up to 300mA was recorded over hundreds of miles/kilometres of telegraph lines in 1891.

Telluric currents can be utilised in mineral exploration, to help locate areas of changes in the electrical conductivity of rocks that may indicate mineral deposits.

## Gravity batteries

A gravity battery is a type of electromechanical battery where a mass is raised and then lowered by gravity to generate electricity. It can be used as a type of energy storage, powering a motor to raise the mass when power is cheap (excess is available) and then lowering it to generate power when it is more expensive (when demand is higher).

Silicon Chip magazine noted some examples of this in their article on Grid-scale Energy Storage (April 2020; siliconchip.au/Article/13801).

A gravity-powered light called the GravityLight was developed for use in less developed countries (see Fig.19). It is 'charged' by raising a 10kg mass by 1m and provides light for five minutes by delivering 20mA continuously. Unfortunately, the project was not a success.

## Hydroelectricity for camping

A portable hydroelectric generator was produced for bushwalkers or campers (Fig.20). You have to anticipate being in an area with reasonably fast-running water. That is not always possible in the Australian bush but is more realistic in parts of the USA, Europe or New Zealand. The device is a bit heavy for many bushwalkers, at 1.5kg, and appears to be no longer available.

## Electromagnetic fields around power lines

It used to be a classic demonstration to hold a fluorescent tube under a high-voltage power line. An electrical current is induced due to capacitive coupling, causing the gases in the tube to fluoresce (see Fig.21). The electric field around a high-voltage power line is shown in Fig.23.

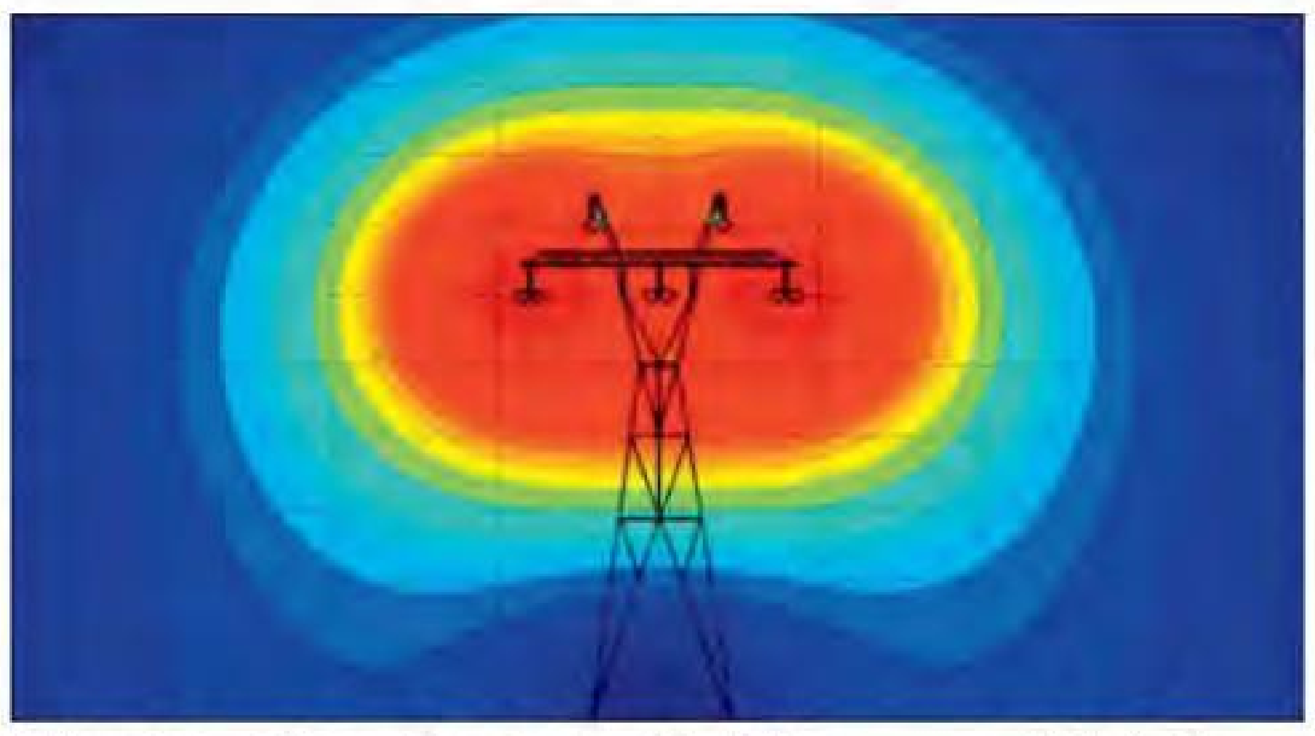


**Fig.23: the electric field around high-voltage power lines. The red region is a reading of >15kV/m. *Source: Quora – https://pemag.com/au/link/abxu***

There must be a sufficient voltage differential between both ends of the tube for it to light. There is a sufficient electrical field gradient to cause the tube to glow if held vertically but not horizontally.

There are many anecdotal accounts (but few documented cases) from the USA of farmers and others building large coils or fences beneath power lines running across their properties to harvest power via electromagnetic induction. It is theoretically possible, but power theft is still illegal even when done 'over the air'.

Very large structures would be required to obtain useful amounts of power (to do more than, say, power some LEDs). With the cost of copper these days, the cost of the wire would exceed any worthwhile savings in electricity, despite the high cost of power. It would be cheaper to buy some solar panels and batteries.

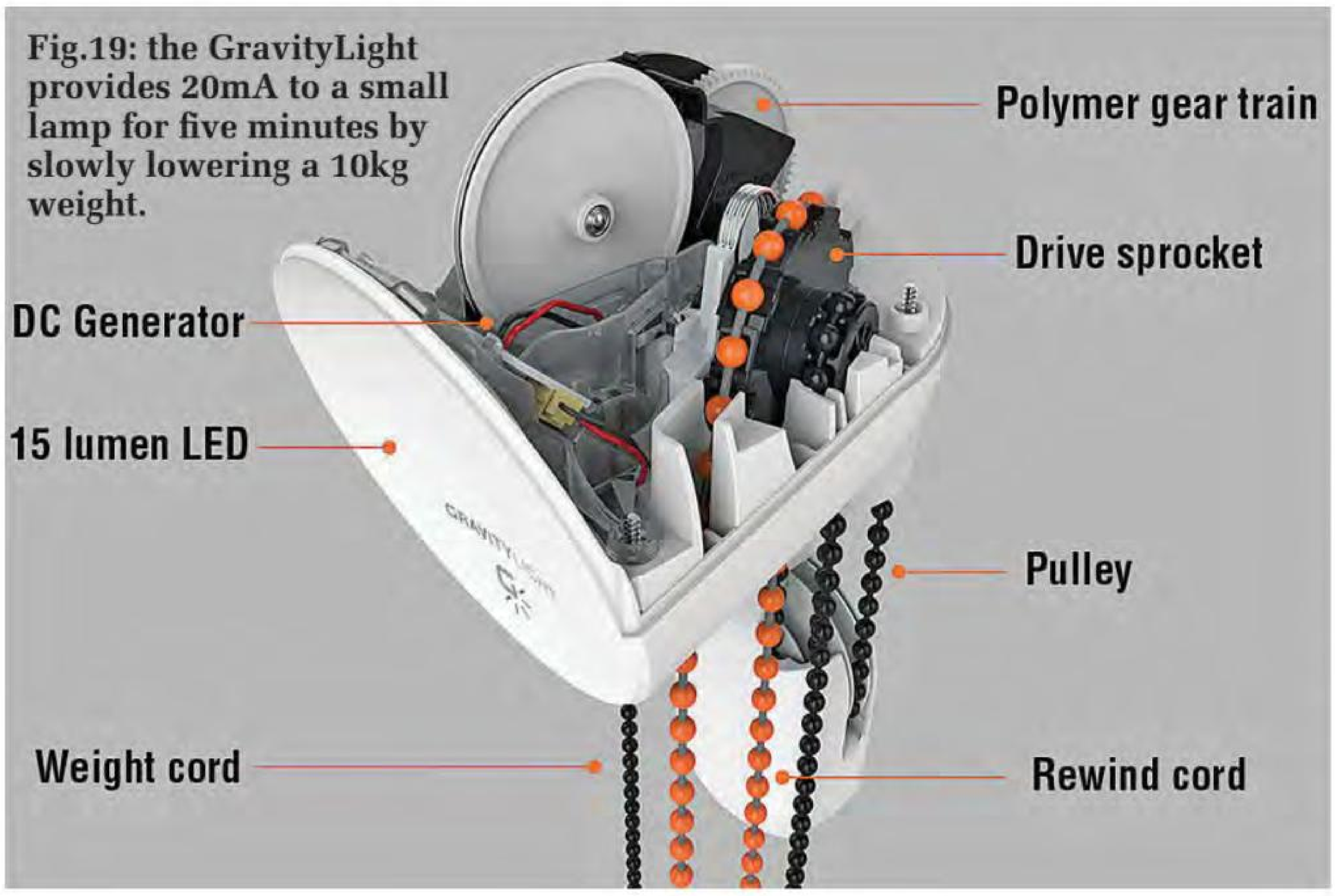


**Fig.19: the GravityLight provides 20mA to a small lamp for five minutes by slowly lowering a 10kg weight.**

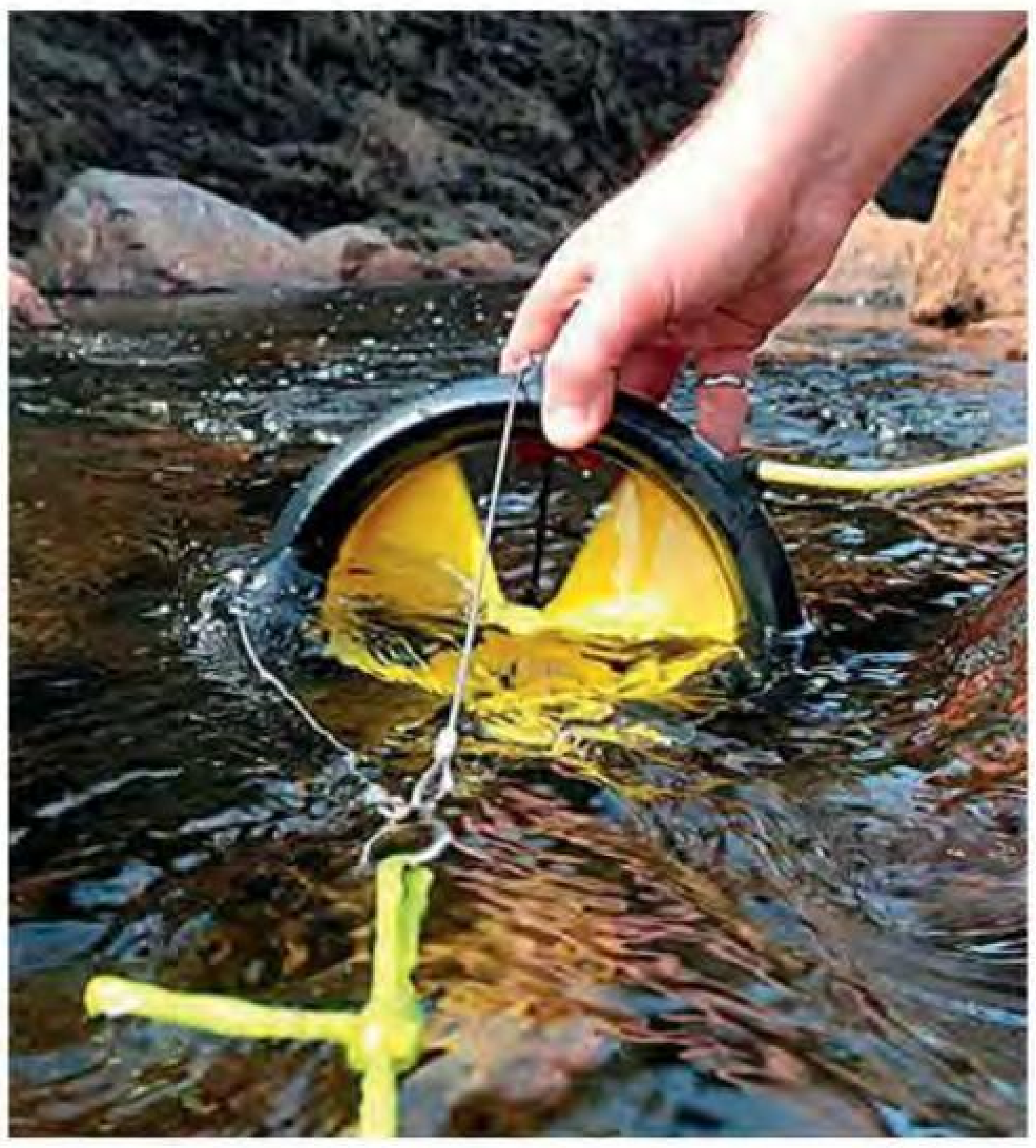


**Fig.20: the "WaterLily Turbine", a portable hydroelectric generator for charging USB or 12V devices in a running stream.**

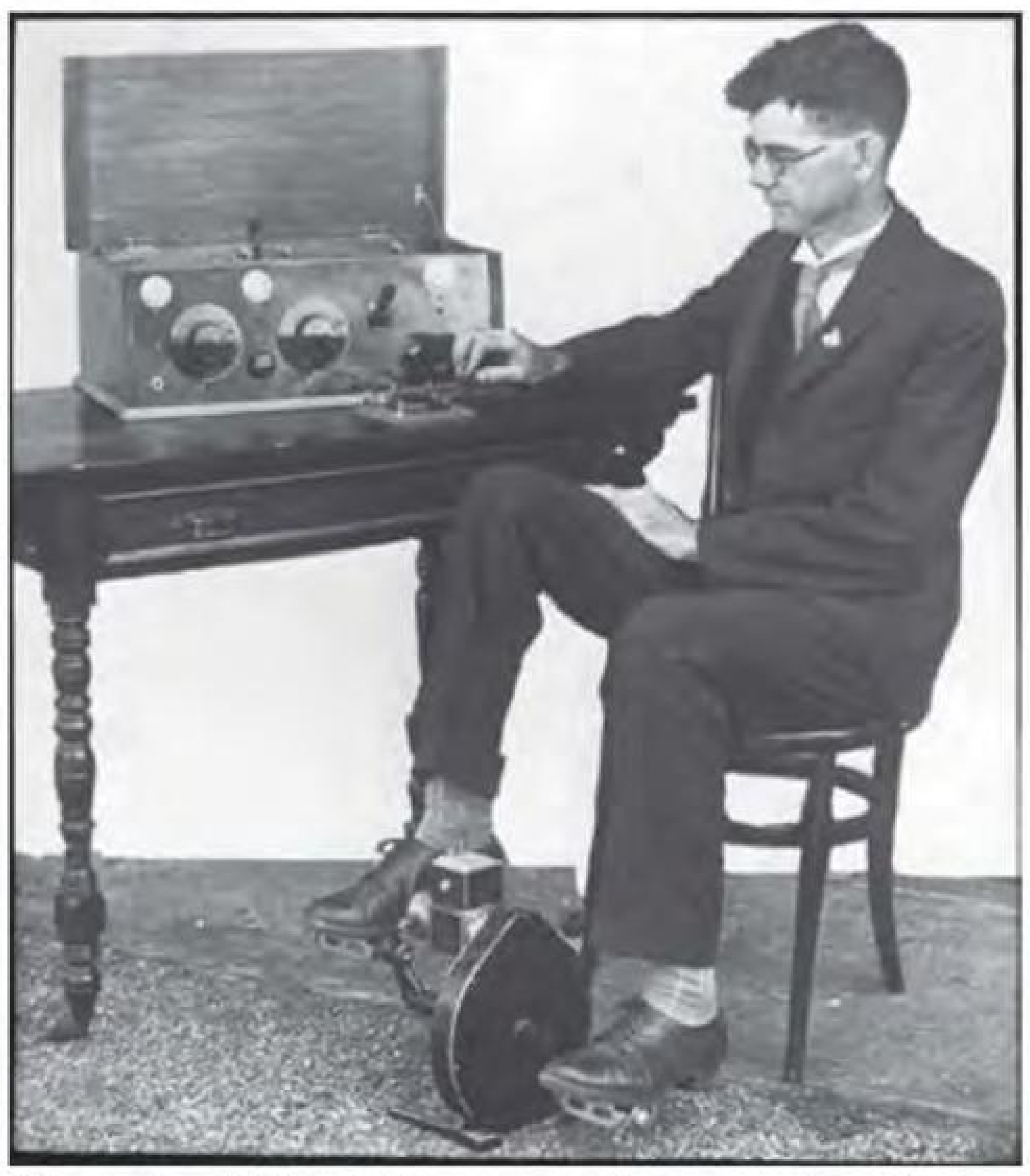


**Fig.24: Alfred Traeger demonstrating the pedal-powered radio he invented in 1928.** ***Source:*** **https://w.wiki/AUk2**

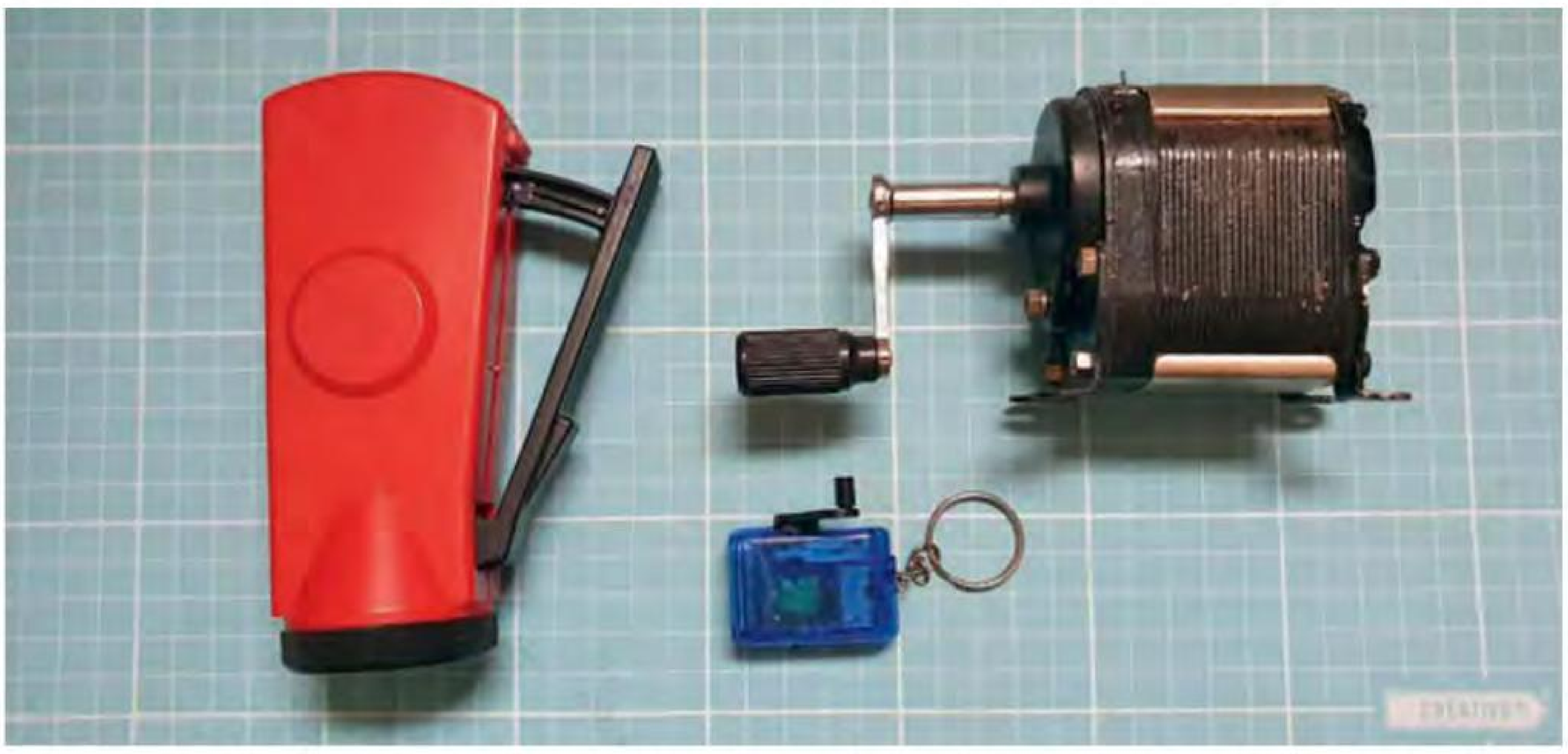


**Fig.25: the Author's collection of hand-cranked devices. The red hand-cranked torch is from the former Soviet Union and has an incandescent bulb, while the blue one is a modern Chinese torch with LEDs and a reserve battery. The item at upper right is a magneto from an old telephone.**

There is an interesting video that explains how to use a coil and capacitor to make a resonant LC circuit to harvest enough power to light an LED from various sources. It is titled "Stealing Electricity (the safe way)" and is at **https://youtu.be/CLS8pbDNHbk**

Also see the video titled "Fences sucking power from under HV transmission lines" at **https://youtu.be/lDm00Ww6qE4**

## Human-powered generators

While pedal-powered generators are less common today due to the low power consumption of LED lights and the advent of lithium-ion batteries, they used to be a common way to power bicycle headlights. They draw power from the rider's pedalling (see Fig.22). They could be either wheel-mounted ('bottle dynamos') or hub-mounted.

They can generate about 3W at 6V (500mA), with some delivering 6W at 12V (also 500mA). Modern hub dynamos such as those from SON can also be used to recharge batteries or mobile devices.

In earlier times, electricity was not readily available in the Australian Outback, so Alfred Traeger invented a pedal-powered radio that was used for the School of the Air and for calling the Royal Flying Doctor Service (see Fig.24). The pedal generator produced around 200V at 100mA (20W). Transceivers from the Traeger Transceivers company were sold to Nigeria in 1962 and Canada in 1970. For further information about Traeger Transceivers visit **https://pemag.au/link/abxf**

A human on a stationary bicycle can drive a higher-power generator, such as to charge a laptop. Instructions to do this are at **https://pemag.au/link/abx8**

There is a large variety of hand-cranked devices that generate electricity for lighting or other purposes, such as those shown in Fig.25.

Many early telephones had a hand crank magneto that generated 50-100V AC to ring a bell at the called party's end, or alert an operator. While current for talking was supplied by batteries, they did not have sufficient power to ring the bell.

Dynamite plungers were similar, although they are now obsolete. They comprised a T-handle attached to a linear rack gear that engaged with a circular gear connected to a generator. When the handle was pressed down, they generated a brief electrical current to trigger a detonator.

## Electric shoes

Experimental shoes have been designed to harvest energy for a variety of possible purposes; one example

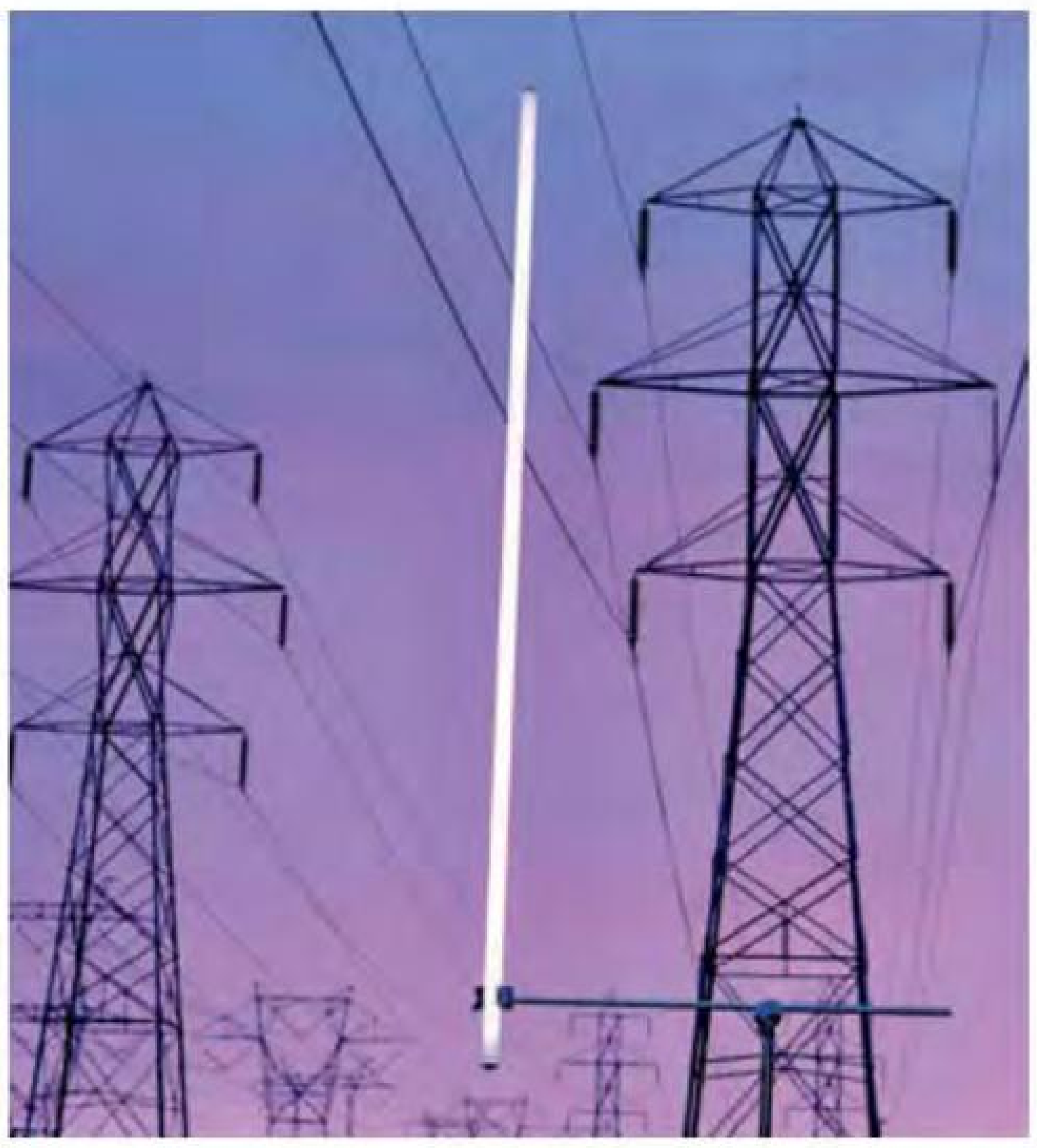


**Fig.21: a fluorescent tube glowing under a high-voltage power line due to capacitive coupling of the electric field.**

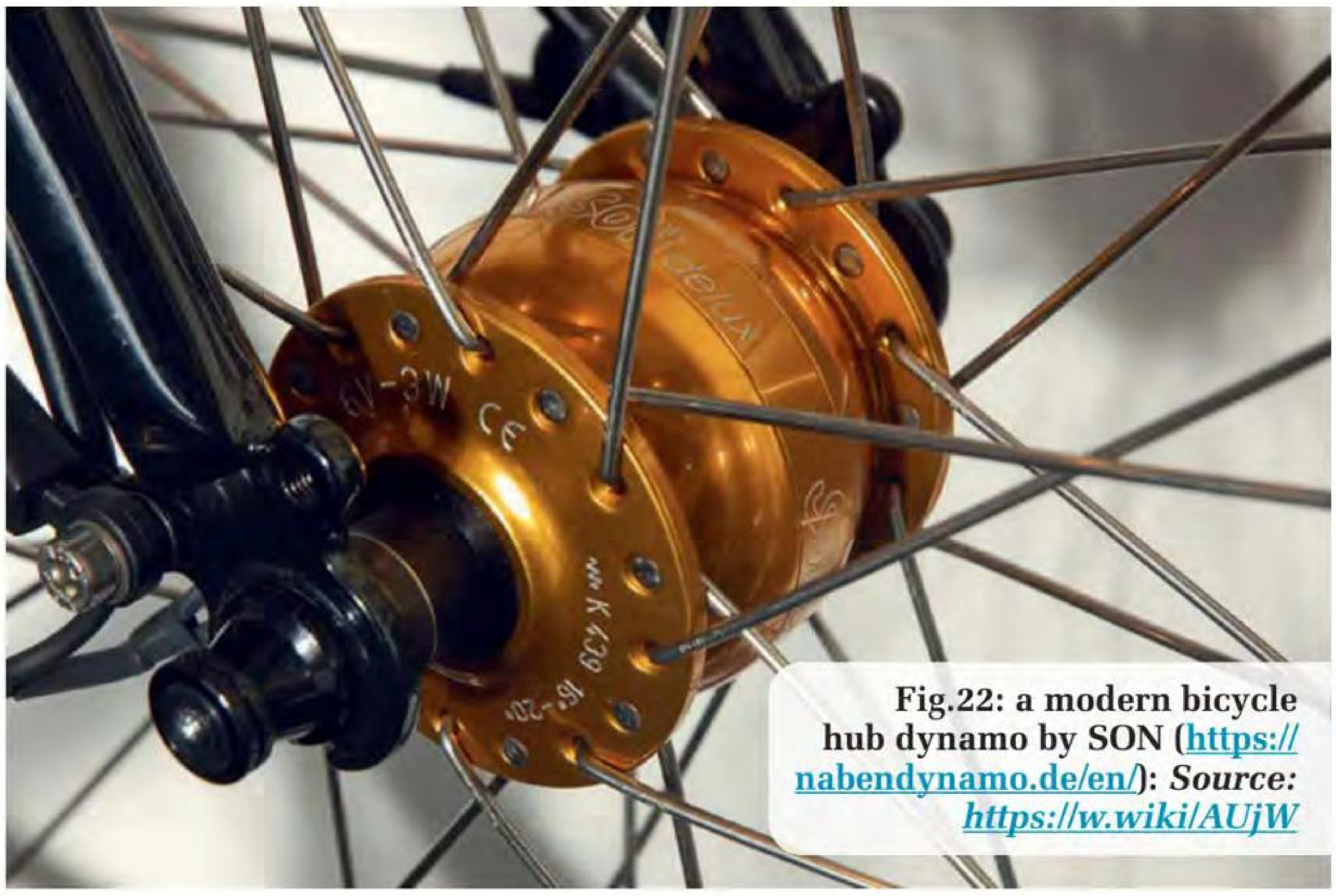


**Fig.22: a modern bicycle hub dynamo by SON (https://nabendynamo.de/en/):** ***Source:*** ***https://w.wiki/AUjW***

is shown in Fig.26. That energy-harvesting combat boot produces electrical power via compression of bulbs in the sole of the boot, which drive microturbines to produce electricity to power a GPS tracker.

Children's shoes that light up generally have batteries and are not self-powered, as explained in the video titled "How Light Up Shoes Work – See What's Inside Sketchers Kids Litebeams" at **https://youtu.be/IIlpRgVBDYo**

On the other hand, kids' scooter wheels that light up do use a small generator built into the hub.

## Power from trains coming down mountains

On page 79 of the April 1988 issue of Silicon Chip magazine, we described how regenerative braking by heavy ore- and coal-laden trains descending the Blue Mountains in Sydney (from mines in places like Lithgow) generated a significant amount of power, which was used to power passenger and empty freight trains ascending into the mountains at the same time.

If ore or other heavy material is mined from mountains and carried down to sea level by trains, which then ascend empty, you effectively have a generator powered by the potential energy of that ore (see Fig.31).

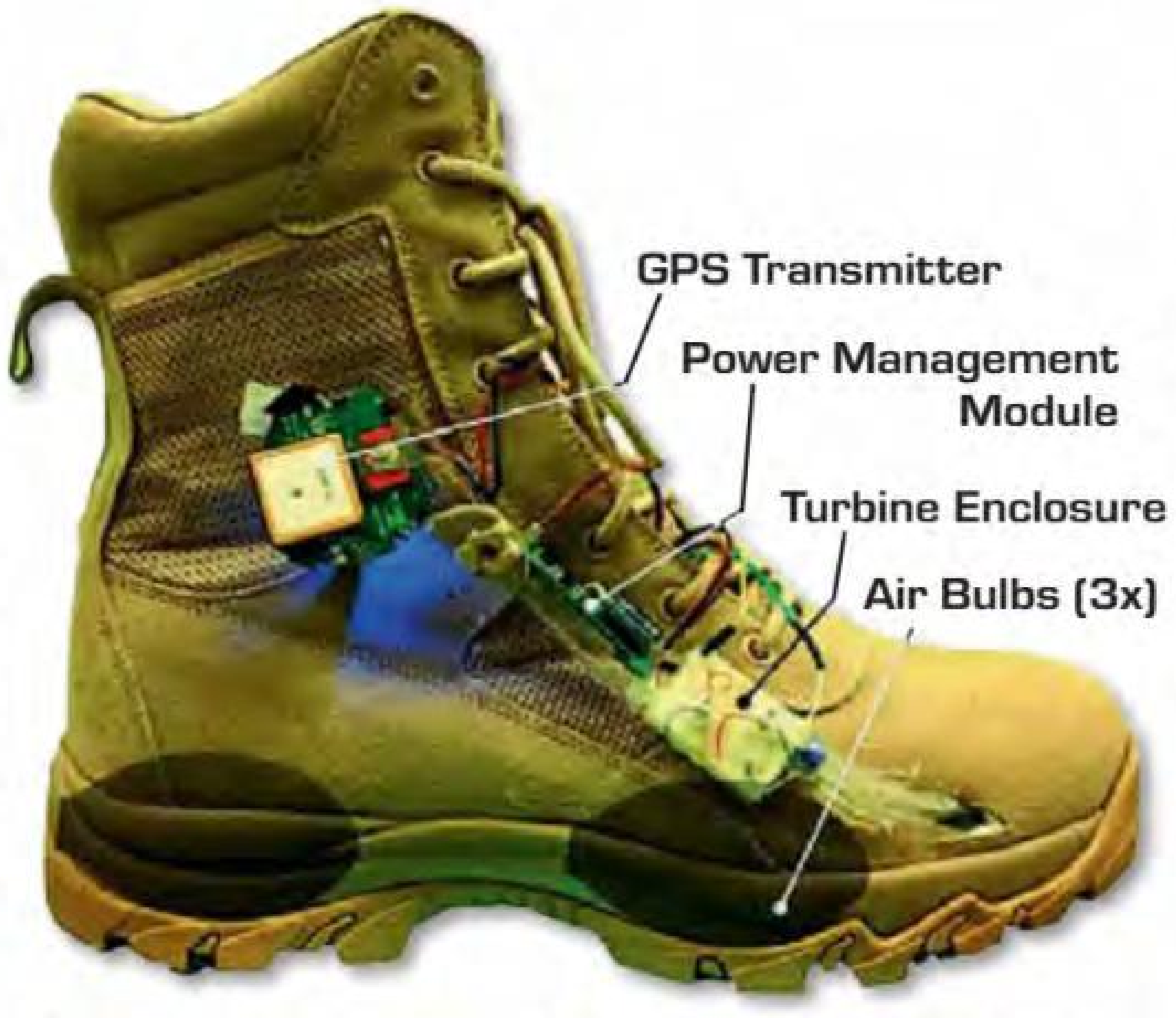

GPS Transmitter
Power Management Module
Turbine Enclosure
Air Bulbs (3x)

**Fig.26: an energy-harvesting combat boot that powers a GPS tracker.** ***Source: www.researchgate.net/publication/325211019***

Fortescue is developing an iron ore freight train in Australia that will charge batteries as it coasts down hills, to provide power for the return journey uphill to get more ore.

## Power from roads

Energy-harvesting experiments have been performed for roadways. Methodologies that have been tried, shown in Fig.27, include:

- Harvesting thermal differentials in between pavement and lower levels underground.
- Devices exploiting Faraday's Law of Induction to harvest mechanical energy (a magnetic field will interact with an electric circuit to produce an electromotive force).
- Piezoelectric devices to harvest mechanical energy.
- Solar panels embedded in, around and above roadways.

## Electrodynamic tethers

An electrodynamic tether is a long wire deployed from an Earth-orbiting spacecraft – see Fig.28. As it passes through the Earth's magnetic field, a current flow develops, according to Faraday's Law of Induction. It can be used as a power source, but it results in some drag on the spacecraft.

In 1996, NASA deployed a long tether from the Space Shuttle Columbia, which generated a potential of 3500V. The tether was intended to be 20.7km long but an electric arc caused the tether to break after 19.7km had been spooled out.

It works as follows – ionospheric electrons are collected from the positively-biased anode at the end of the uninsulated tether. They flow through the electrical load, then to the negatively-biased cathode, where they are discharged into

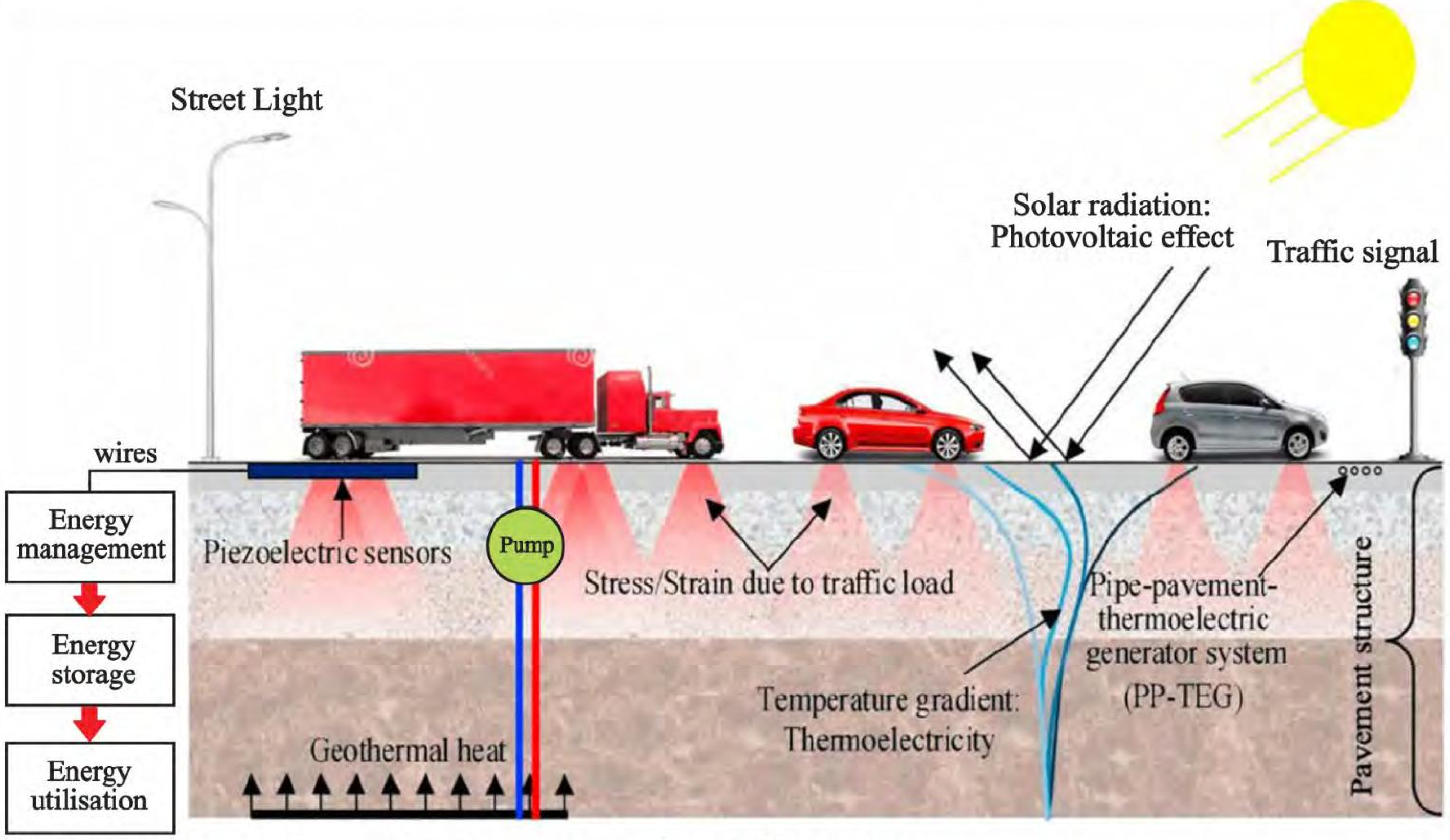

Street Light
wires
Energy management
Energy storage
Energy utilisation
Piezoelectric sensors
Pump
Geothermal heat
Stress/Strain due to traffic load
Solar radiation: Photovoltaic effect
Traffic signal
Pipe-pavement-thermoelectric generator system (PP-TEG)
Temperature gradient: Thermoelectricity
Pavement structure

**Fig.27: some concepts of energy harvesting from vehicles travelling on roads.** ***Original source: www.mdpi.com/1996-1073/16/7/3016***

the space plasma and complete the circuit.

## Electrostatic generation from lunar soil

NASA has proposed harvesting the electrostatic charge from lunar soil. The charge builds up over long periods due to the solar wind. They propose to collect the charge using a moving capacitor array that's 'raked' through the lunar soil (see **https://pemag.au/link/abxj**).

NASA estimates that a ⅓m² collecting array could produce a maximum theoretical power of 147W (700V @ 0.21A) – see Figs.29 & 30.

## Tiny solar cells

Inexpensive, tiny solar cells can be used to power IoT or sensor devices, with energy stored in a small battery or cell. Even photodiodes can be pressed into service to generate power; see Fig.32.

People in the developed world might not appreciate it, but for people living in less developed countries, nighttime lighting is not always available and it is highly beneficial if they can get it. Certain charities, such as SolarAid (**https://solar-aid.org**), produce solar lights for people in these countries, and donors can also buy them for their own use.

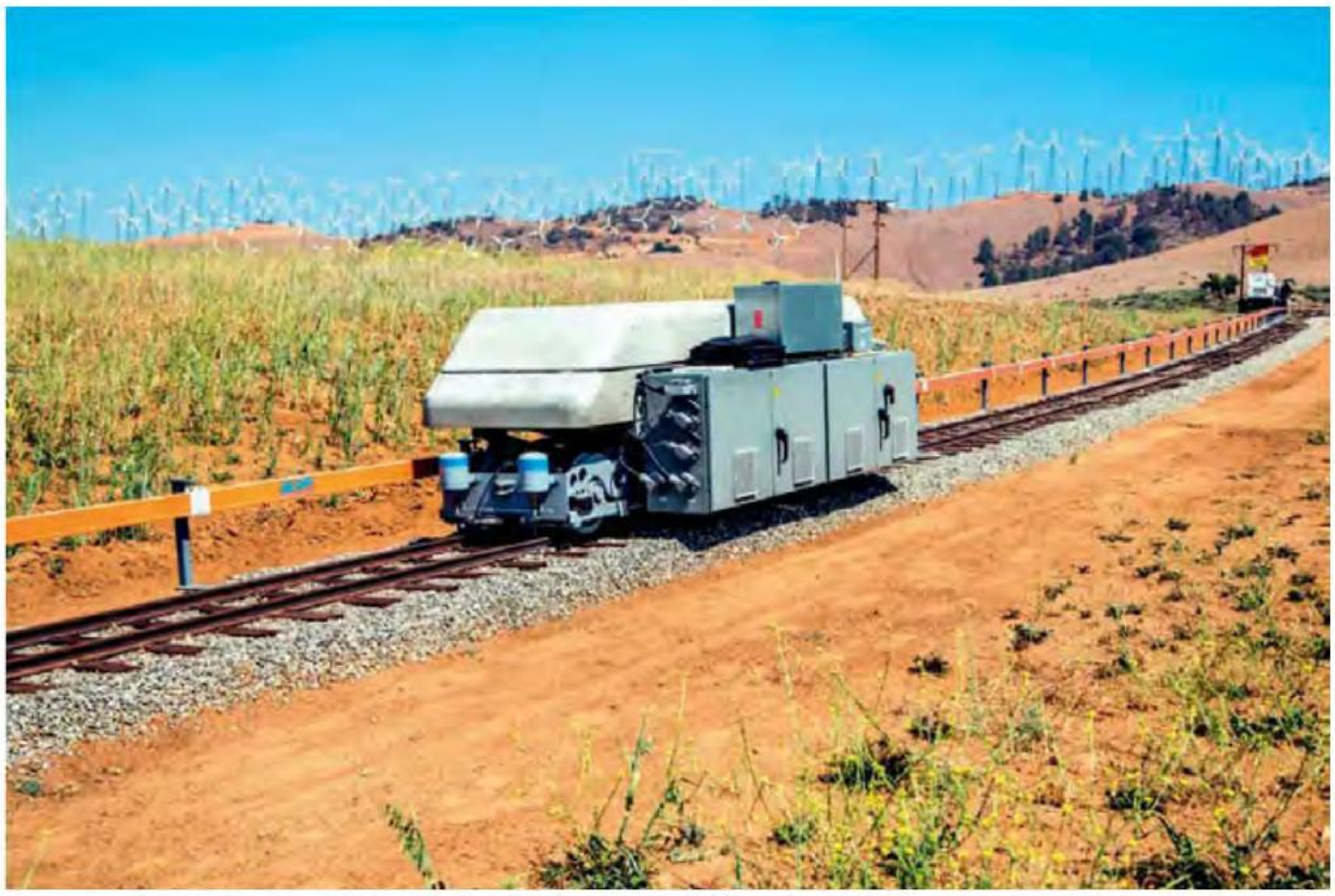


**Fig.31: the ARES rail car, which climbs a hill using electricity during off-peak hours, then is released downhill during peak hours to produce energy via regenerative braking.** ***Source: ARES North America – aresnorthamerica.com***

Many small solar panels are available for ramblers and campers to recharge devices. Some can be affixed to backpacks, while others are set up when camped. However, panels that are small and light enough to be affixed to a backpack provide only small amounts of power. I find that you typically get to a campsite well after peak sun. In my experience, it is better to carry batteries.

## Micro hydroelectric schemes

New Zealand YouTuber Marty T made a 'microhydro' installation on his wilderness property using the motor from a scrap Fisher & Paykel Smart-

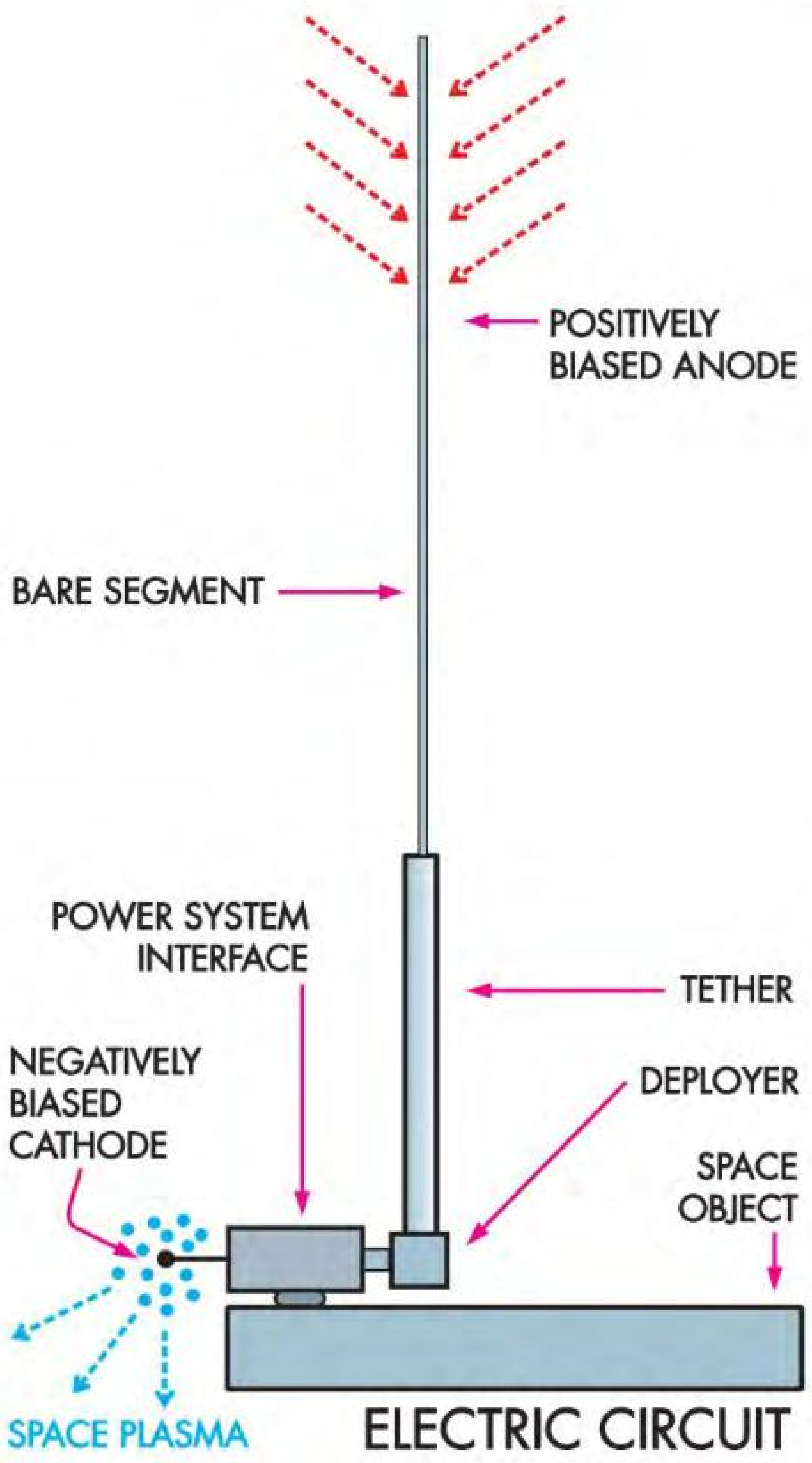

POSITIVELY BIASED ANODE
BARE SEGMENT
POWER SYSTEM INTERFACE
TETHER
NEGATIVELY BIASED CATHODE
DEPLOYER
SPACE OBJECT
SPACE PLASMA
ELECTRIC CIRCUIT

**Fig.28: an electrodynamic tether deployed from a spacecraft.** ***Original source: https://w.wiki/AUjv***

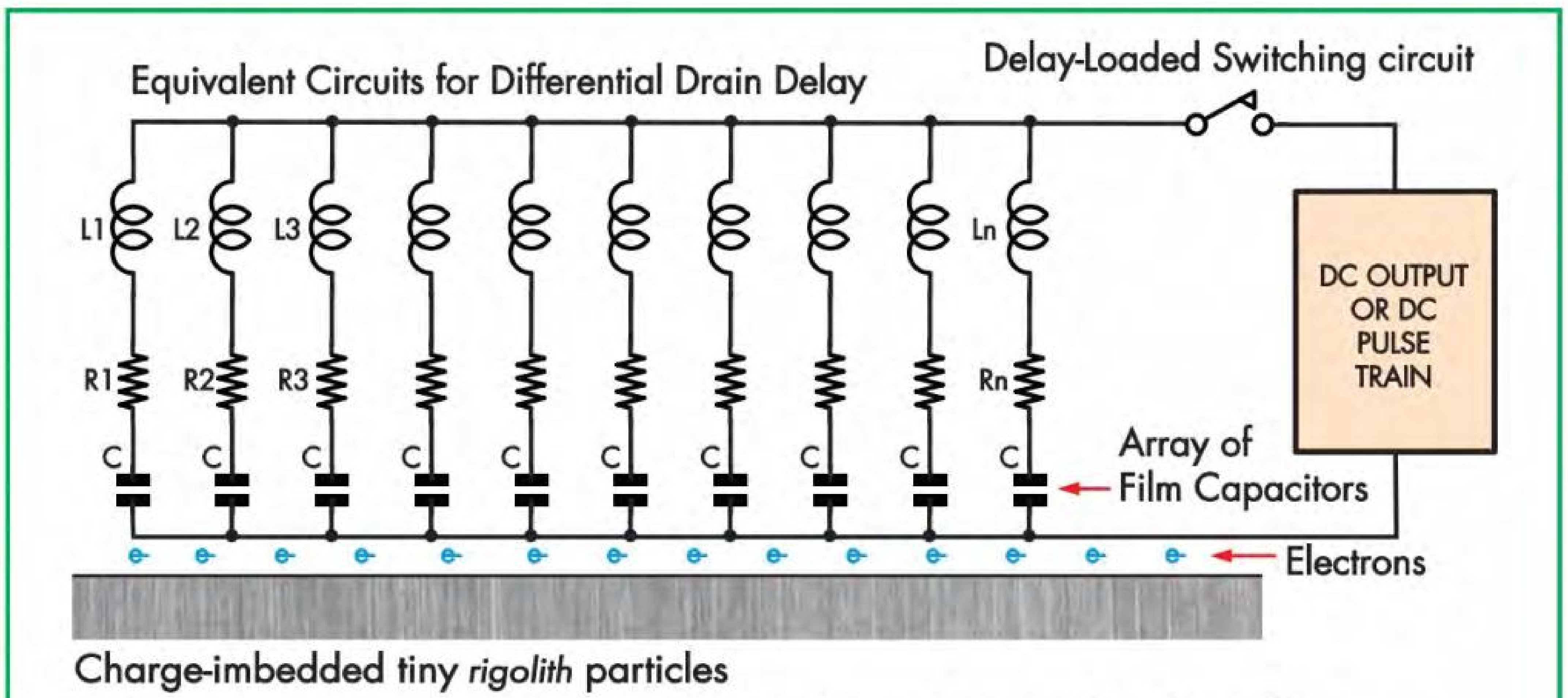

Equivalent Circuits for Differential Drain Delay
Delay-Loaded Switching circuit
L1
L2
L3
Ln
R1
R2
R3
Rn
C
DC OUTPUT OR DC PULSE TRAIN
Array of Film Capacitors
e
Electrons
Charge-imbedded tiny rigolith particles

**Fig.29: the circuit of a theoretical capacitive charge collector with a differential drain to harvest electrostatic charge from the negatively charged lunar soil (regolith).** ***Original source: https://ntrs.nasa.gov/api/citations/20100032922/downloads/20100032922.pdf***

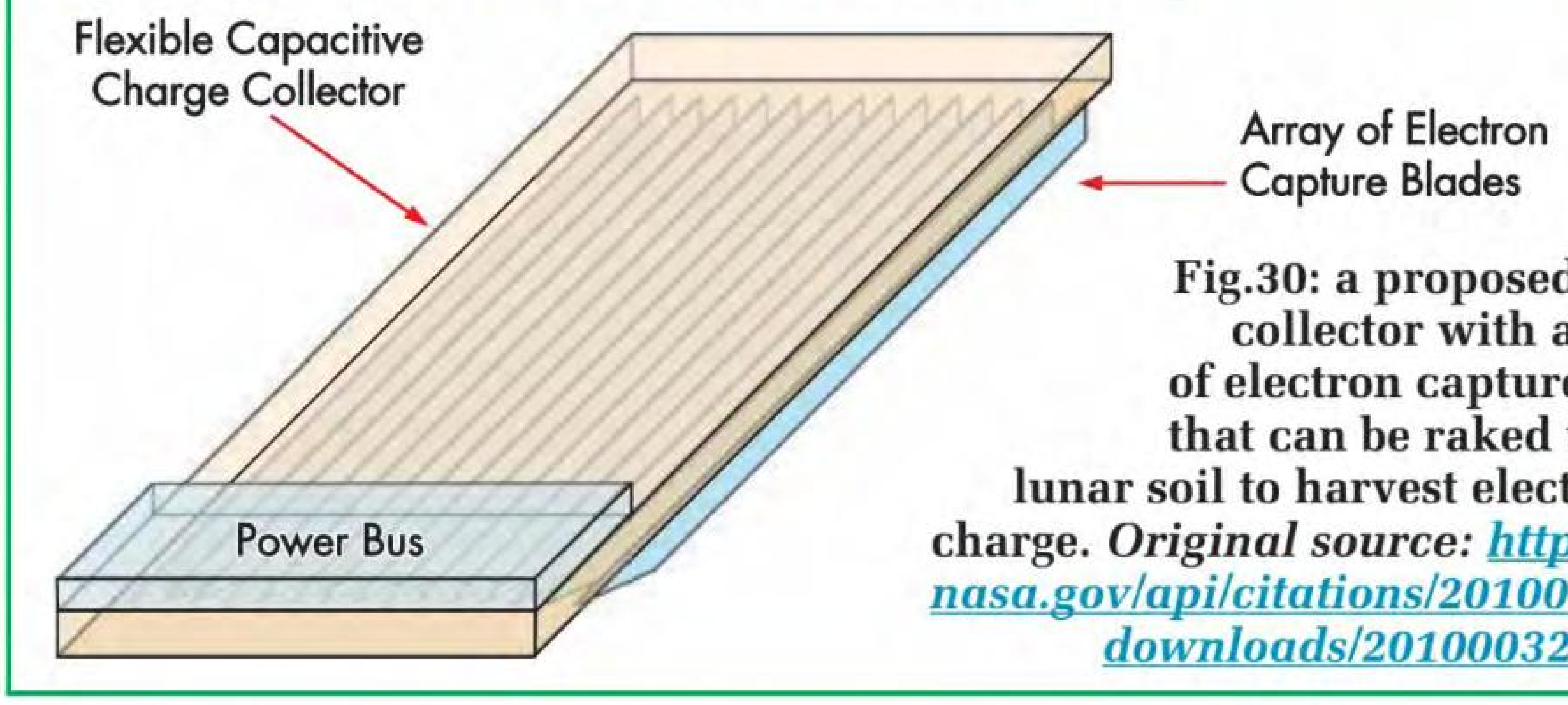

Flexible Capacitive Charge Collector
Array of Electron Capture Blades
Power Bus

**Fig.30: a proposed charge collector with an array of electron capture blades that can be raked through lunar soil to harvest electrostatic charge.** ***Original source: https://ntrs.nasa.gov/api/citations/20100032922/downloads/20100032922.pdf***

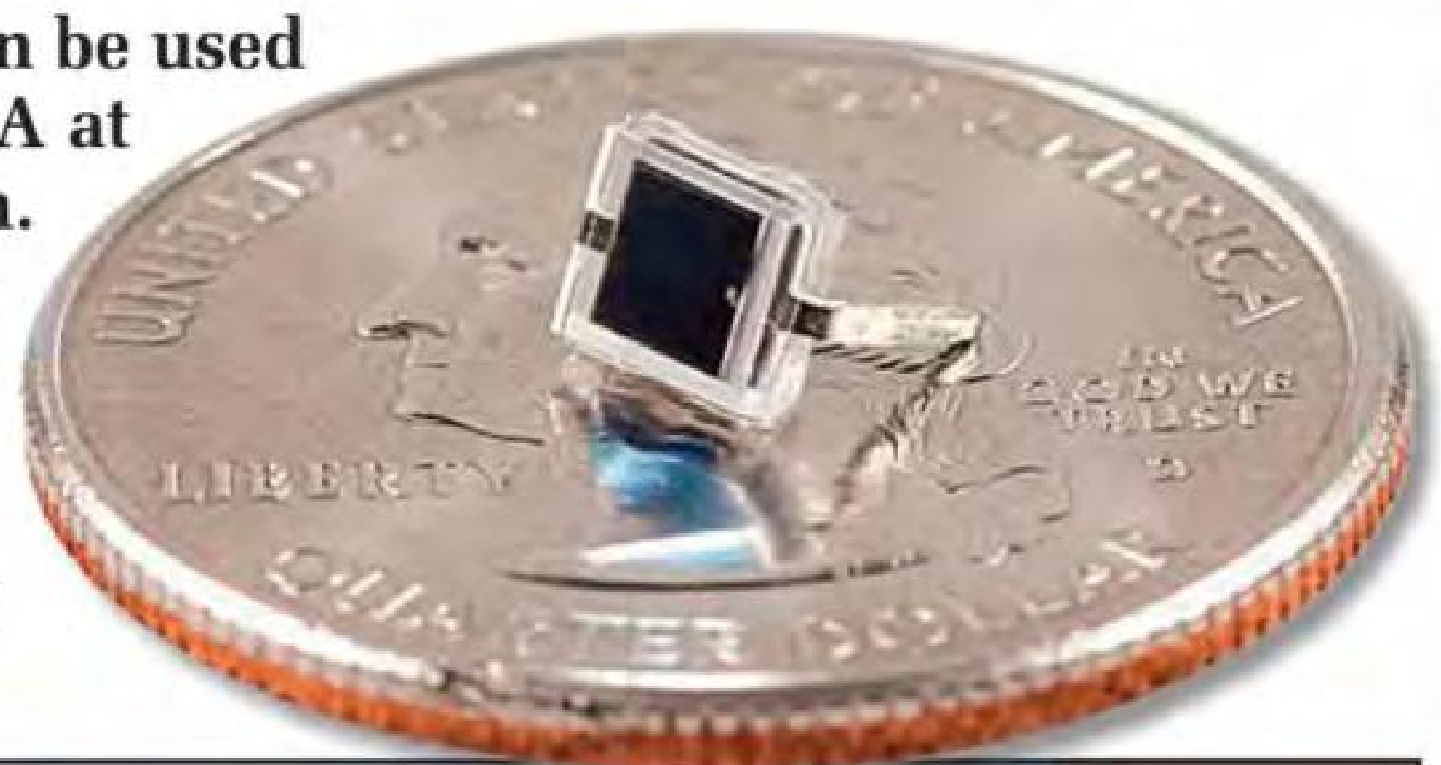
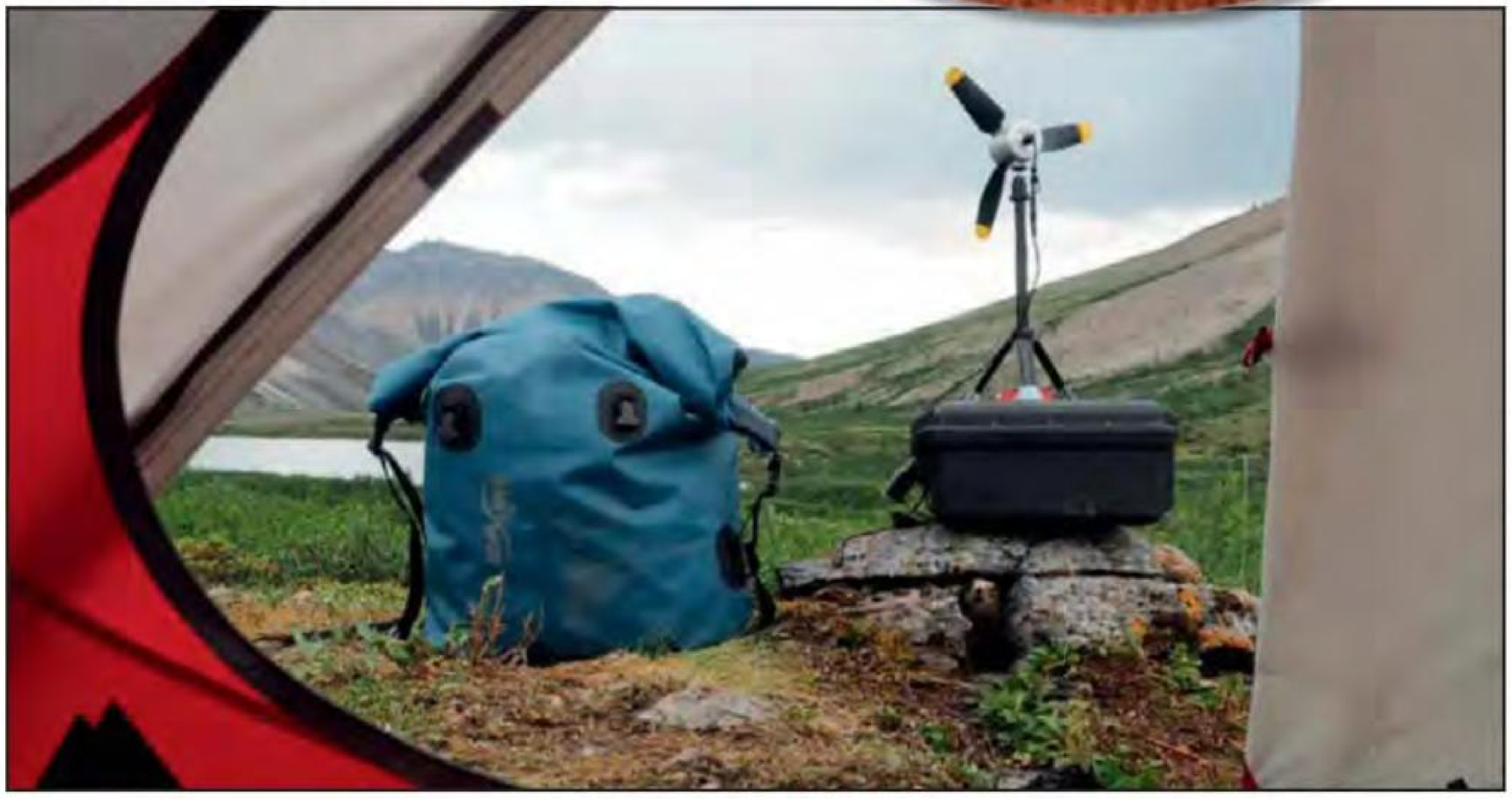


Fig.32: a BPW34 PiN photodiode can be used as a solar cell, producing up to 47µA at 350mV. The coin diameter is 24.26mm. *Source: Core Electronics PRT-09541 – https://pemag.au/link/abxz*

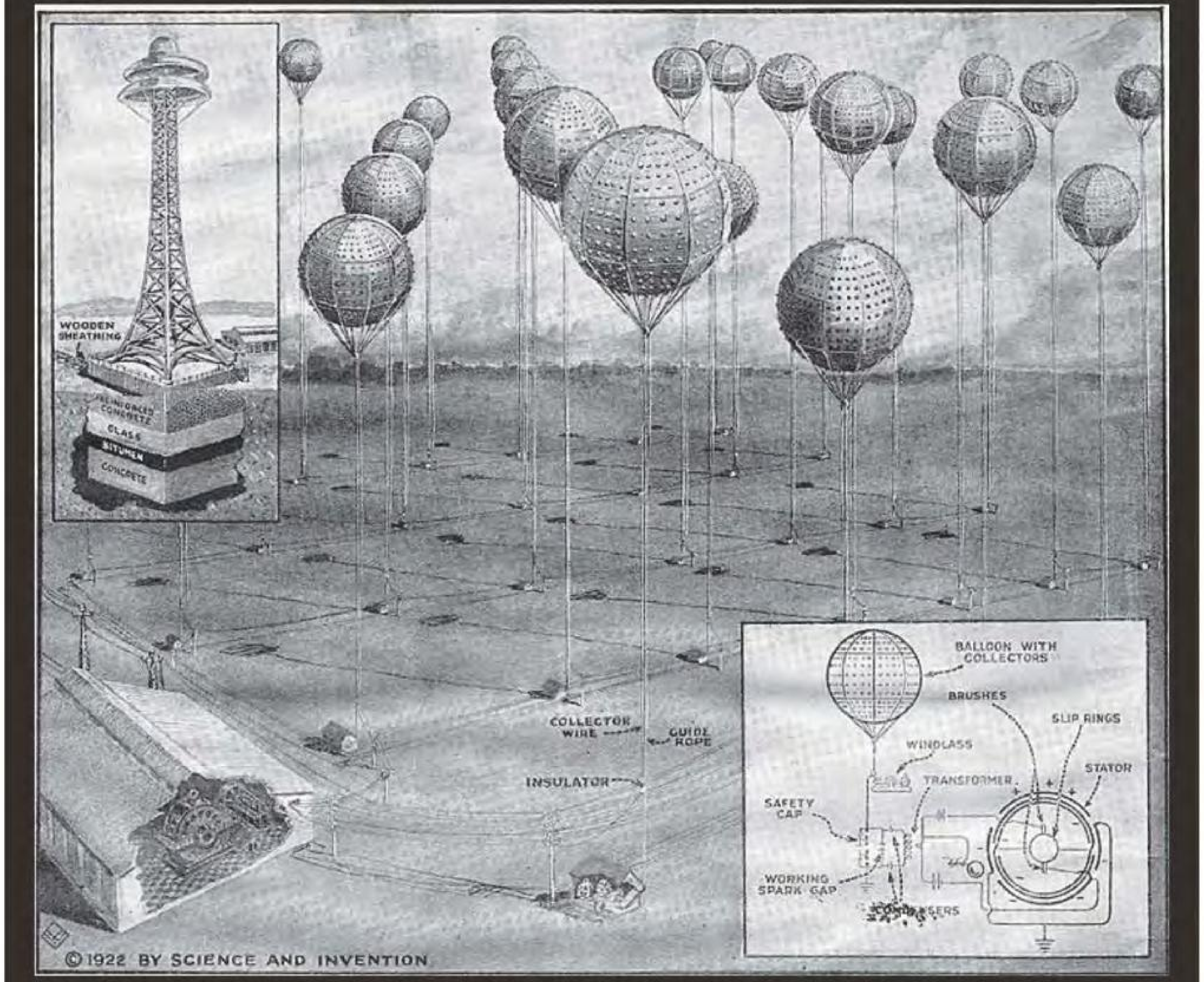
Fig.33 (below): a wind turbine that can be used for camping. *Source: Tex Energy – https://pemag.au/link/abxx*



## Energy harvesting is not new

In 1925, Estonian inventor Hermann Plauson obtained US Patent 1540998 for "Conversion of atmospheric electric energy". He proposed harvesting atmospheric electricity with a network of balloons. H. Gernsback earlier described this idea in "Science and Invention", February 1922 (**https://pemag.au/link/abxg**).

It is unlikely this would have been practical. However, it was claimed in the description that a single balloon at 274m altitude could provide 400V at 1.8A, which certainly would be useful if attained! We suspect that it was under unusual atmospheric conditions and could not be achieved regularly.



***The energy harvesting idea of Hermann Plauson.***
***Source:* www.reddit.com/r/Air_Fountain/comments/1cc3dx6/**

Drive washing machine, converted to a DC generator.

The motor has to be rewired to reduce the voltage and increase the current, to make it more suitable for charging a battery bank. Details of motor rewiring are at https://pemag.au/link/abxk, but many other resources explain how to do it. Also see this series of videos:

1. https://youtu.be/LVoeaKCEd2o
2. https://youtu.be/lbuvTSWh50U
3. https://youtu.be/8SWq5Pskpug

A US YouTuber decided to see if he could make a hydroelectric system powered by rainwater collected on a roof. He calculated that 2W could be generated from rain falling on a house roof and going down the downpipes, but on his first attempt, he only got 0.19W. On his second attempt, he generated over 0.61W and, on the third attempt, over 0.91W. Of course, it has to be raining for this to work.

In Australia, such a system might work best in the tropics, such as Far North Queensland.

See the videos for more details:

1. https://youtu.be/S6oNxckjEiE
2. https://youtu.be/YLb4enCgnP4
3. https://youtu.be/vify0k2sHlQ

## Portable wind generators

A wind generator can be used for bushwalking, provided it is anticipated there will be reasonable wind at the campsite. A model such as the Infinite Air 5T can produce up to 5V at 2A and weighs 1.65kg (Fig.33 shows the larger 3.2kg Infinite Air 18 model). As with the portable hydroelectric generator, we feel the weight is too high for most potential use cases.

## MEPAP

The MEPAP ("Multipurpose and source Electricity Generator with Air Purifier") is something Heath Robinson or Rube Goldberg might have dreamt up.

It harvests electricity using vibration (piezoelectric materials), electromagnetic radiation (metamaterials), electromagnetic induction (inductive coupling), wind energy (mini turbine with dynamo) and thermoelectric energy, all to operate an air purifier device.

It is described at https://pemag.au/link/abxh, but we don't know how well it works. PE

# The Fox Report

Barry Fox's technology column

## The analog loophole is now digital

**The analog loophole is alive** and well, and extremely useful. But it is now seldom mentioned because young executives in the music and movie industries have never heard of it and wouldn't understand it, anyway. The few older execs who once tried, and failed, to plug the hole now understand it will always exist.

All manner of clever technical tricks based on encryption are used to try to control digital copying or 'cloning' of words, graphics, speech, music and movies. They can at least frustrate the process. But if the material is analog – and it always has to end up that way because humans have analog senses – it is simple to copy.

Yes, there are ways to watermark or fingerprint analog sound and pictures with invisible or inaudible codes, but this is used mainly to identify the source. Yes, this can be extended to copy control, but only if the copy system is hard-wired to recognise the codes and block the copy process.

No-one chooses to buy hardware that is designed not to do what's wanted. Blu-ray disc players can impose copy control, but this roadblock has been sidelined by ripping software for PCs with Blu-ray drives.

The main and most frustrating copy control system in real-world use is HDCP (High-bandwidth Digital Content Protection, developed by Intel) which is part of the HDMI (High-Definition Multimedia Interface) cable system which is now universally used for digital sound and vision connection.

When an HDMI cable is connected between a "source", such as a Blu-ray player, DVD player, DVR, computer, game console or music player, and a "sink", such as a TV set, projector or hifi system, the two ends of the HDMI chain 'talk to each other' in an attempt to allow listening or viewing while blocking copying.

Some years ago, even some legitimate connections (for instance, between a TV tuner or game console and video projector) were frustrated because projectors did not have HDMI inputs, just analog 'component' video sockets. This is why Sky put high-definition component outputs on its early digital boxes.

Also, gamers wanted to capture or stream their gameplay. Home moviemakers wanted to do the same with their own creations.

Enterprising companies like Hauppauge then designed boxes that converted analog component high-definition (or composite standard-definition) video to USB digital video that could be digitally captured by a PC running free software such as OBS (for video) or Audacity (for audio). This created an analog loophole.

Then, without any fanfare, conversion dongles from China slid onto internet sale sites; these little units, the size of a USB memory stick, converted HDMI to USB digitally. Most importantly, the Chinese dongles simply ignored HDCP. And they cost only around £20-30, quickly falling to the current price of around £10.

Plug one end of the dongle into any HDMI output socket and the other end into a PC running OBS or Audacity (or similar video capture software) and what you get is a digital recording or stream of whatever the HDMI socket is outputting, be it gameplay, music or movie material.

This isn't even an analog loophole; it's an easy and convenient digital loophole because there is no need for any digital-analog-digital conversion; just an HDMI digital signal converted directly to a USB digital signal.

It has always been a mystery why the HDMI licensing authorities, music and movie industries let this happen. Perhaps they thought the dongles were hard-to-find and sold only in small quantities by under-the-radar companies. Perhaps they never got their heads around the clever technical trick the Chinese were playing with these absurdly inexpensive dongles.

I've been experimenting with HDMI-to-USB dongles for years. Forget about perish-the-thought uses like copying TV recordings from a PVR; these dongles are a wonderful tool for capturing and storing old audio and video recordings such as amateur recordings and home movies.

Just convert the original analog to HDMI using one of the many readily available, low-cost analog-to-digital units and then use the low-cost HDMI-to-USB dongle and a PC to capture PCM audio and high-definition video.

Now, finally there is what I believe to be the first HDMI-to-USB converter dongle available from a 'legitimate' Western source. It's made in China, of course – what now isn't? – but it carries the highly respectable Sandberg name (which I have previously mentioned in connection with other electronics and IT gadgetry).

Sandberg's brief instruction guide shows how to connect the dongle to a PC or Mac running OBS (obsproject.com) or Debut (nchsoftware.com). Debut is new to me, so I can't vouch for it with the confidence I have in OBS, after years of using OBS.

Sandberg says the dongle can also be used for live-streaming with online video chat software such as Zoom or MS Teams, or uploading to YouTube, with a video camera serving as a webcam.

The physical connections on the dongle are standard HDMI with USB 3.0 C and A. Input quality up to 4K 30 FPS is handled with capture output up to 1080p. The unit self-powers from the USB connection. No software drivers are needed.

Rather confusingly, the dongle shows up on a PC as a "USB 3.0 audio" device. The official technical specifications from Sandberg reads:

- Input video: 720p/1080p up to 60 FPS, 4K up to 30 FPS
- Input maximum resolution: 4K@30Hz
- Output maximum resolution (minimum USB 3.0): 1080p @ 60Hz
- Supports UVC standard
- 24/30/36-bit deep colour
- Captured video size: maximum 1080p, 60 FPS
- Chipset: MS2130

Says Martin Hollerup, CEO at Sandberg A/S: "It's a missing link. I've personally been on a film shoot in a forest with a laptop and thought it would have been smart if I could just use the laptop's screen as a field monitor during the shoot. I couldn't do that back then, but with this new adapter, it would have been super easy."

The only mention in the documentation that may raise an eyebrow with the entertainment industry is: "You can also load your favourite DVDs onto your laptop before a flight". But as almost no-one buys DVDs any more (the charity shops are bursting with them for peanut prices), this is hardly likely to ring alarm bells in Hollywood.

It's the other unmentioned uses of all these capture dongles that would clang bells, but my bet is that only a handful of managers in Hollywood have any idea what a capture dongle is or does and how it creates a barn-door-wide loophole.

The Sandberg price is a bit higher (around £30, compared to around £20 plus postage for a no-name dongle), but with the Sandberg name, you get a five-year warranty and support.

# Variable Speed Drive Mk2

## For Induction Motors

### Part 2 by Andrew Levido

**Last month, we introduced the Mk2 VSD and described its features, circuit and firmware. This month, we cover construction, testing and some hints for using it.**

SPEED
STOP
RUN
REVERSE

**Everything, including the heatsink and fan,** is mounted on a single printed circuit board (PCB) that fits into an ABS plastic enclosure measuring 220 × 165 × 60mm, as shown in the accompanying photographs.

Many of the components are surface-mount types, but they are all relatively easy to solder by hand. There are no fine-pitch chips, and the passives are all 2 × 1.2mm or larger, except for three diodes, which are a little bit smaller but should be manageable. Anyone with a modicum of SMT soldering experience should have no trouble putting it together successfully.

That said, this is a complex build, and because of the high voltages and currents involved, it is recommended only for experienced constructors.

Regardless of your skill level, if you build this, you must follow the safety instructions when it comes to the testing stage. It's also a good idea to double and triple-check your work before powering it up. We'd hate for you to put a lot of effort into building this, only for it to blow up because something was installed backwards or in the wrong spot.

## Assembly

We recommend assembling the VSD in two stages, as described below. First, we will focus on the control circuitry, so we can test it safely at a low voltage and get it working. After that, we will move on to the power electronics.

The VSD is built on a double-sided board coded P9048-C or 11111241 that measures 150 × 205mm. Start by fitting all the surface-mounting parts, using the overlay diagram (Fig.8) and close-up of the section near the microcontroller (Fig.9). Work methodically across the board, paying attention to the orientation of polarised components like ICs, diodes (including LEDs) and electrolytic capacitors. You can also refer to the silkscreening on the PCB.

We won't go into a great amount of detail here on how to solder SMD parts, as it is now pretty common, and many of our projects require it. However, we'll give a quick overview and some tips. There are three main ways you could solder the SMDs: with a reflow oven, with a hot air rework station or with a soldering pencil/iron.

Those with reflow ovens and hot air rework stations likely are already familiar with the required techniques, which involve adding solder paste to the board, placing the components on top and then heating the solder paste until it reflows.

Manual soldering is best done with a syringe of good-quality flux paste. For each part, spread a thin layer of flux paste on the pads, then place the part on its pads, ensuring it is correctly orientated. One of the worst things you can do is solder an IC to the board backwards! For the microcontroller in a quad flat package, there are four possible orientations, but only one is correct (with the pin 1 dot as shown).

With the part in place and a clean soldering iron, add a little solder to the tip and tack-solder one of the part's pads. Check that all its pins are lined up with the other pads; if not, the joint can be remelted and the part gently nudged into position. Once in position, the remaining pins can be soldered and the initial one refreshed.

Finally, for parts with closely spaced pins (like ICs), check for solder bridges between pins. If found, they can be cleared with the application of a little more flux paste and then solder-wicking braid. The braid can also be used to remove excess solder if there's too much on some pins.

Once all the surface-mounting parts are in place, clean the flux residue off the board, then add relay RLY2, DIP switch bank S1, trimpots VR1 & VR2, header CON17 and the input terminal blocks, CON8-CON11. Slot all four blocks together (in dovetail fashion) before soldering them in place.

# Variable-Speed Drive Mk2, part two

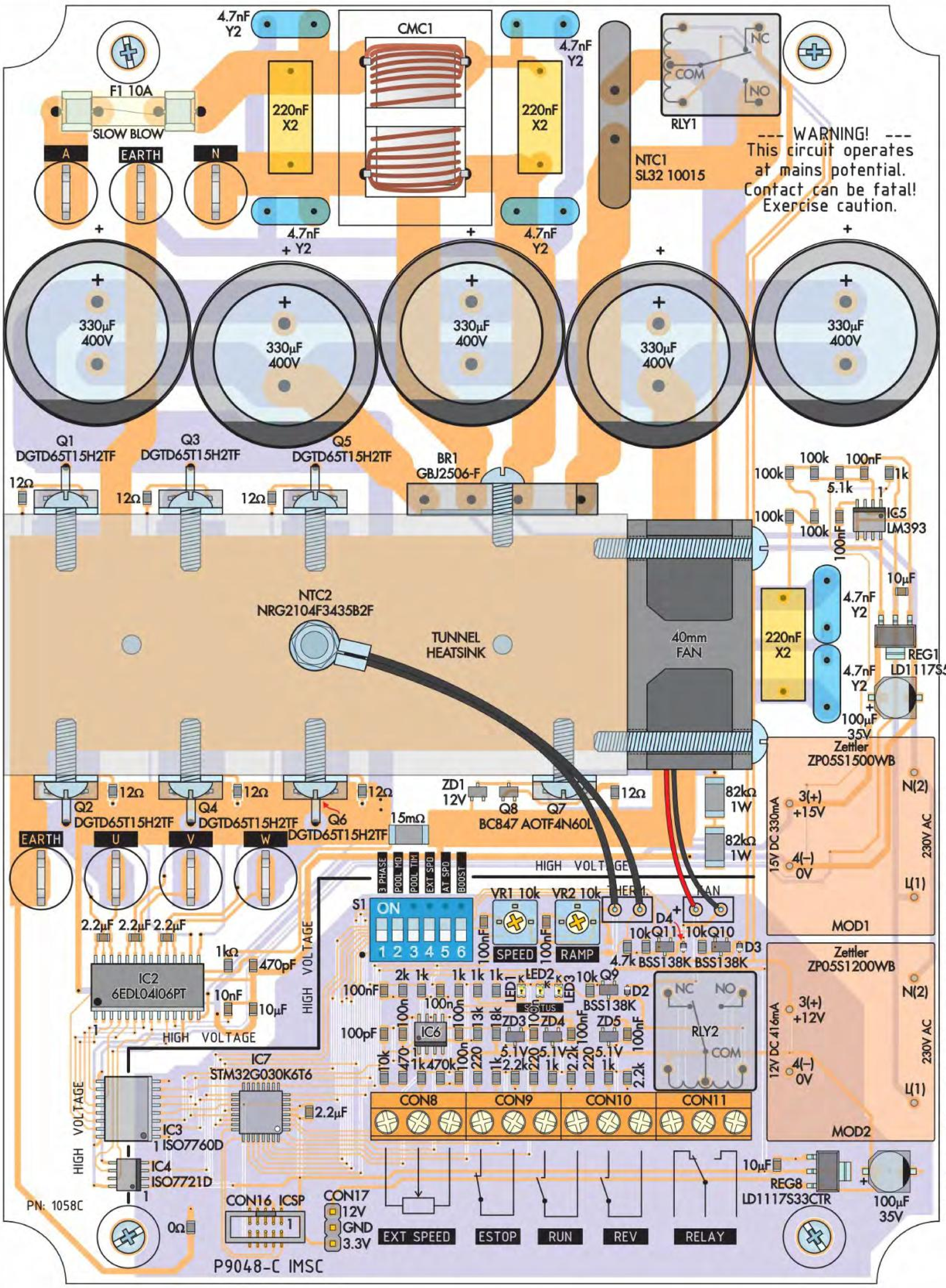

4.7nF Y2
CMC1
220nF X2
F1 10A
SLOW BLOW
A
EARTH
N
RLY1
NC
COM
NO
NTC1
SL32 10015
--- WARNING! ---
This circuit operates
at mains potential.
Contact can be fatal!
Exercise caution.
330µF 400V
Q1 DGTD65T15H2TF
Q3 DGTD65T15H2TF
Q5 DGTD65T15H2TF
BR1 GBJ2506-F
12Ω
100k
100nF
5.1k
1k
IC5 LM393
NTC2
NRG2104F3435B2F
TUNNEL HEATSINK
40mm FAN
10µF
REG1 LD1117S50
100µF 35V
Zettler
ZP05S1500WB
3(+) +15V
4(–) 0V
15V DC 330mA
230V AC
N(2)
L(1)
MOD1
Q2 DGTD65T15H2TF
Q4 DGTD65T15H2TF
Q6 DGTD65T15H2TF
ZD1 12V
Q8 BC847
Q7 AOTF4N60L
15mΩ
82kΩ 1W
EARTH
U
V
W
HIGH VOLTAGE
S1
3 PHASE
POOL MD
POOL TIM
EXT SPD
AT SPD
BOOST
ON
1 2 3 4 5 6
VR1 10k
VR2 10k
SPEED
RAMP
THERM.
FAN
D4
10k
Q11
Q10
BSS138K
4.7k
D3
2.2µF
IC2 6EDL04I06PT
1kΩ
470pF
10nF
10µF
100nF
100pF
2k
LED1
LED2
LED3
STATUS
Q9
D2
ZD3 5.1V
ZD4 5.1V
ZD5 5.1V
IC6
13k
18k
470k
2.2k
220
RLY2
Zettler
ZP05S1200WB
3(+) +12V
4(–) 0V
12V DC 416mA
MOD2
IC7 STM32G030K6T6
IC3 ISO7760D
IC4 ISO7721D
CON8
CON9
CON10
CON11
REG8 LD1117S33CTR
100µF 35V
PN: 1058C
CON16 ICSP
CON17
12V
GND
3.3V
0Ω
EXT SPEED
ESTOP
RUN
REV
RELAY
P9048-C IMSC

**Fig.8: this component overlay shows where everything goes on the PCB. Fit the surface-mounting parts first, then the DIP switch, trimpots and relay RLY2. Test the control circuitry thoroughly, as described in the text, before moving on to the power electronics.**

## WARNING: DANGEROUS VOLTAGES

This circuit is directly connected to the 230V AC mains. As such, most of the parts and wiring operate at mains potential. **Contact with any part of these non-isolated circuit sections could prove fatal.**

Note also that the circuit can remain potentially lethal even after the 230V AC mains supply has been disconnected! To ensure safety, this circuit MUST NOT be operated unless it is fully enclosed in a plastic case. Do not connect this device to the mains with the lid of the case removed. **Do not touch any part of the circuit for at least 30 second after unplugging the power cord from the mains socket.**

*Do not attempt to build this project unless you understand what you are doing and are experienced working with high-voltage circuits.*

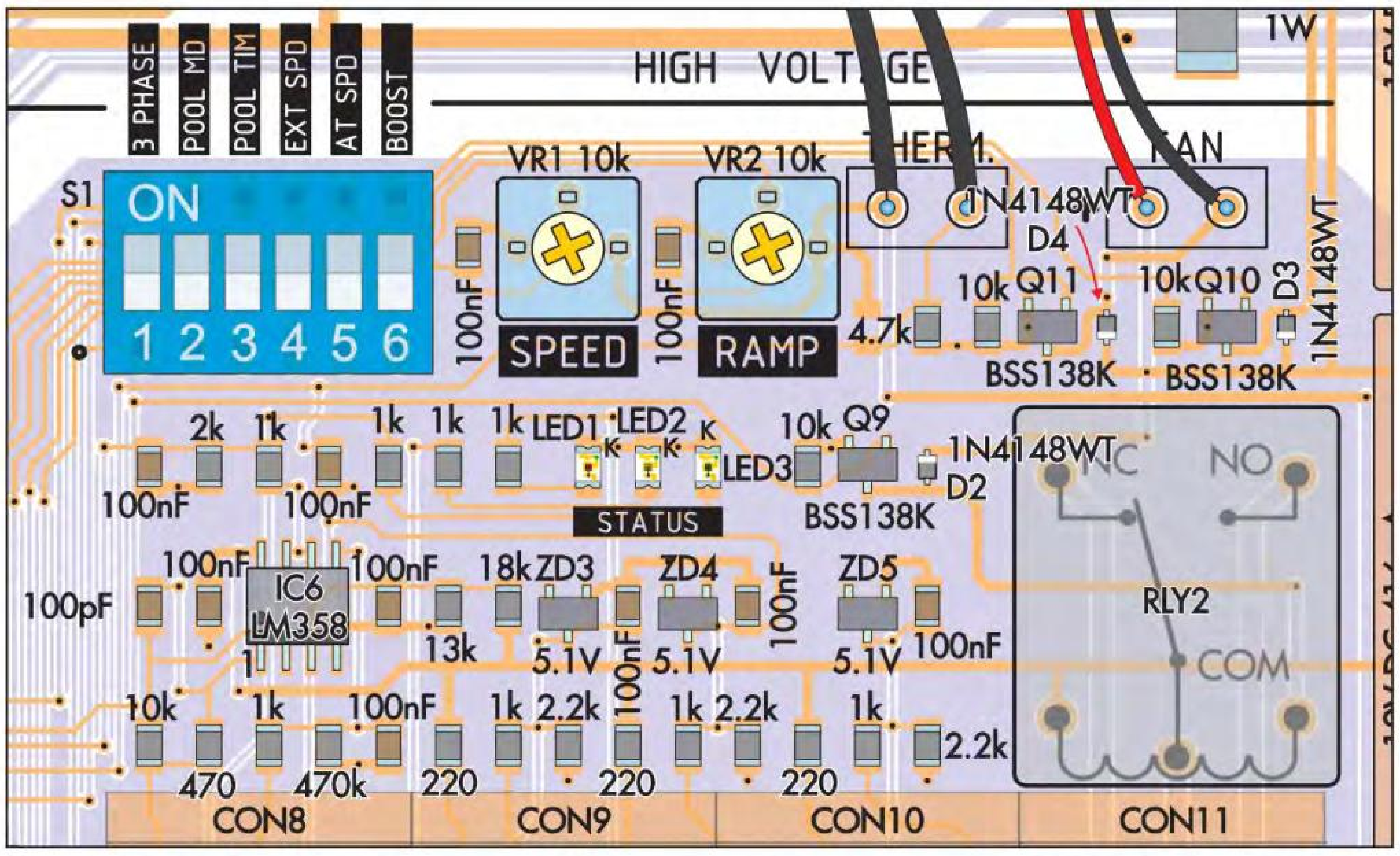
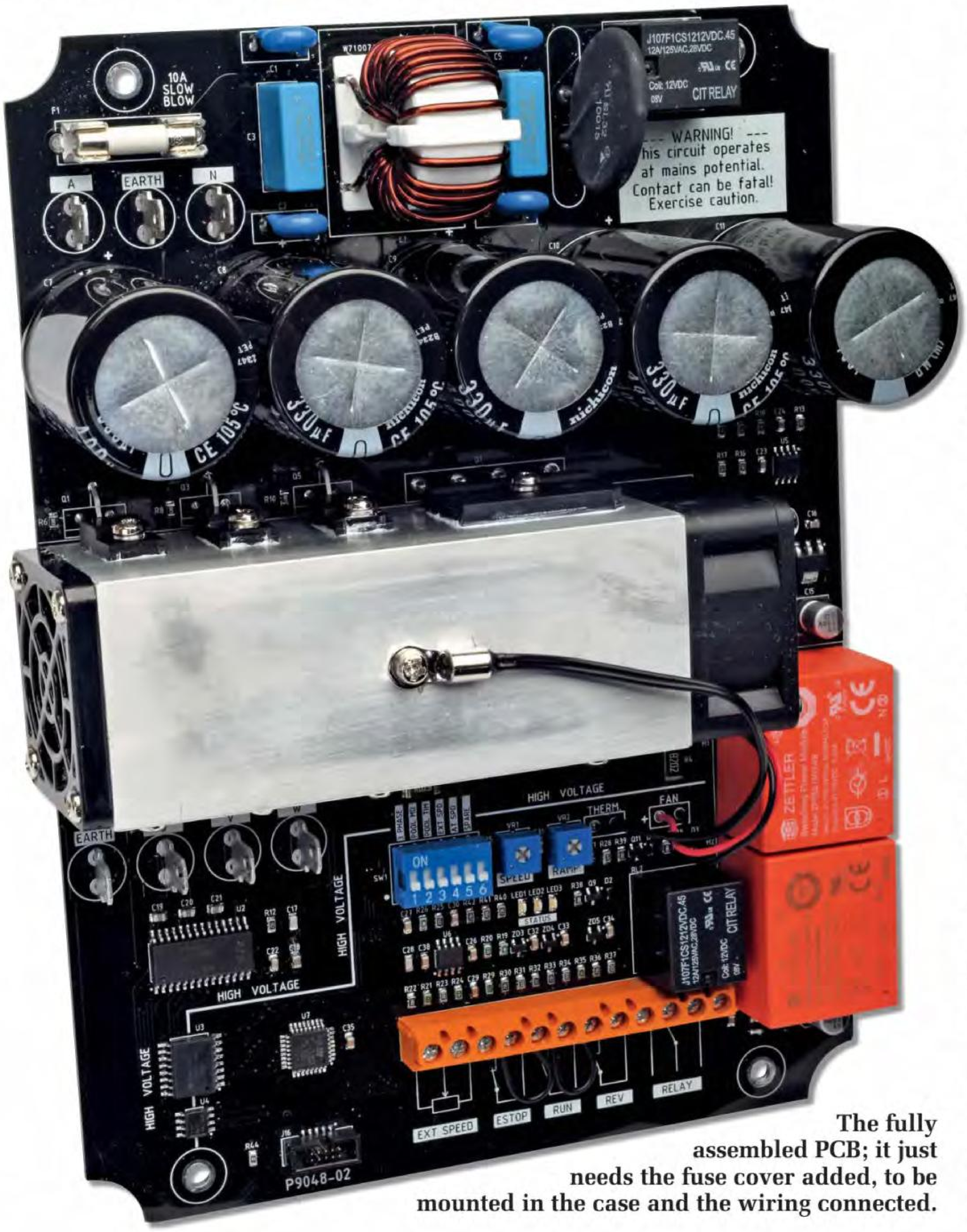

3 PHASE
POOL MD
POOL TIM
EXT SPD
AT SPD
BOOST
HIGH VOLTAGE
1W
S1
ON
1 2 3 4 5 6
100nF
VR1 10k
SPEED
VR2 10k
RAMP
THERM.
FAN
1N4148WT
D4
10k Q11
10k Q10
D3
1N4148WT
4.7k
BSS138K
BSS138K
2k
1k
1k
1k
1k
LED1
LED2
LED3
K
10k Q9
1N4148WT
D2
BSS138K
STATUS
NC
NO
100nF
100nF
100nF
IC6
LM358
100nF
18k ZD3
ZD4
ZD5
100pF
13k
5.1V
5.1V
5.1V
100nF
RLY2
COM
10k
1k
100nF
1k 2.2k
1k 2.2k
1k
2.2k
470
470k
220
220
220
CON8
CON9
CON10
CON11

**Fig.9: this close-up of Fig.8 shows the most densely populated section, so that you can more clearly see the values of the resistors and capacitors there.**


10A SLOW BLOW
F1
A
EARTH
N
WARNING! This circuit operates at mains potential. Contact can be fatal! Exercise caution.
CIT RELAY
330µF
nichicon
CE 105°C
ZETTLER
HIGH VOLTAGE
THERM
FAN
ON
1 2 3 4 5 6
SPEED
RAMP
STATUS
EARTH
U
V
W
HIGH VOLTAGE
EXT SPEED
ESTOP
RUN
REV
RELAY
P9048-02

**The fully assembled PCB; it just needs the fuse cover added, to be mounted in the case and the wiring connected.**

At this stage, you will have installed all the parts in the low-voltage domain except for the AC-to-DC switch-mode power supply module, MOD2. We can now test this circuitry.

Connect a bench supply to the +12V and GND pins of CON17. Make sure the polarity is correct and don't accidentally connect it to the +3.3V pin! DuPont jumper leads are a good way to make this connection.

Set the supply to deliver 12V DC, with the current limit set at around 200mA. When you switch it on, the power supply should not go into current limiting. If is does, there is a short circuit or incorrectly placed component somewhere, so switch off and check the components on the board carefully, including their solder joints.

## Initial testing

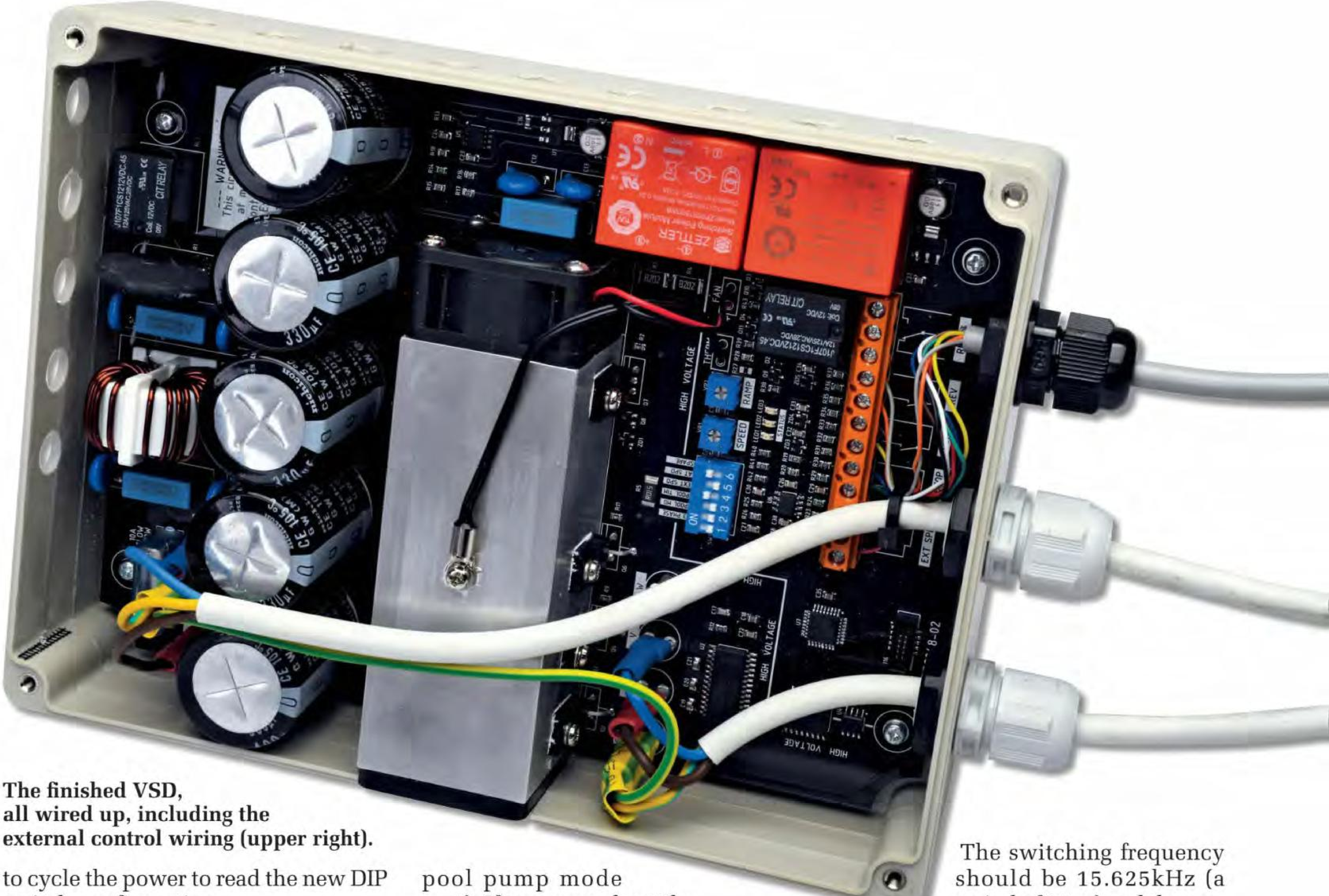
If the current draw is OK, check for 3.3V at the bottom pin of CON17 relative to GND. It should be in the range of 3.1-3.5V. If that is OK, and your microcontroller is not already pre-programmed, now is the time to connect an ST-Link programmer to CON16 and flash the code using the STM32Cube software (a free download). If yours is pre-programmed, you can skip this step.

With the micro programmed, to check for the correct operation of the control circuit, first ensure all the DIP switches are in the off positions and both trimpots are wound all the way anticlockwise, then apply power. All three LEDs should flash briefly twice, then after about three seconds, the yellow LED (LED2) should come on.

If you short the E-Stop & Run pairs of terminals with two wire links and advance the speed trimpot (VR1), the yellow LED should extinguish and the green LED (LED3) should flash while the speed ramps up to the setpoint, at which time LED3 will light steadily.

If you turn the speed pot back to zero, the controller should ramp down with the green LED flashing until the yellow LED lights again. Increasing the ramp time using trimpot VR2 should prolong the ramp time.

If you close the At-Speed DIP switch and repeat the above process, you should hear relay RLY2 close whenever the green LED stops flashing and lights steadily, then open when it begins to flash again. Don't forget that you need



**The finished VSD, all wired up, including the external control wiring (upper right).**

to cycle the power to read the new DIP switch configuration.

You can try opening the Run switch or the E-Stop circuits while the speed controller is running (green LED on or flashing). If Run is opened, the green LED should flash while the speed ramps down to zero, then the yellow LED should light. If the E-Stop switch is opened, the yellow LED should come on immediately.

Now you can check pool pump mode. Bridge the E-Stop and Run terminals again, set the speed and ramp potentiometers to about halfway and close the pool pump mode DIP switch ("POOL MD").

On reapplying power, the controller should start and ramp to full speed with the green LED flashing slowly. After about 30 seconds, the speed should ramp down (green LED flashing fast) to the preset speed (green LED on steadily). Trying again with the Pool-Time DIP switch ("POOL TIM") also closed should extend the pool-pump period to about five minutes.

You can check three-phase mode by closing that DIP switch. It should work as described earlier (ignoring the pool pump mode part). If you now short the Reverse terminals while it is running, the speed should ramp down (green LED flashing fast) then stop for two seconds (yellow LED on) and ramp up again to the preset speed.

Finally, you can check fault operation by momentarily shorting out the thermistor terminals. The red and yellow LEDs should latch on. Opening and reclosing the E-Stop circuit should reset the fault.

If you hit a snag at any point, stop, check the board carefully and fix the problem. Each step above tests a different part of the circuit, so consult the relevant part of the circuit diagram for components to check. Fix any problems and verify it has the correct operation before moving on.

If you have an oscilloscope, you can take a look at the PWM motor drive signals on pins 2 to 7 of IC3. They can be a bit difficult to trigger on since the pulse widths are continuously varying, so consider using one-shot mode to capture a snapshot if your 'scope supports it. There will only be signals on four of these pins if single-phase mode is selected.

The switching frequency should be 15.625kHz (a period of 64µs) and the amplitude about 3.3V.

## Power electronics

Start the assembly of the power components by preparing the heatsink. This is a 100mm length of 40 × 40mm heatsink 'tunnel' extrusion. Mine came cut to length from AliExpress. A total of 11 holes need to be drilled and tapped in accordance with the drilling diagram (Fig.10).

There is a different arrangement of holes on each face, so take care to get them all in the right orientation with respect to each other. I recommend clearly labelling each face according to the diagram and marking the fan end. Mark the hole positions, but before drilling anything, offer it up to the board to check the marks line up with the IGBTs, Mosfet and diode bridge.

Don't forget to run the tap through the four extruded corner 'holes' on each end to make the mounting of the fan and finger guard easier. Use some wet & dry abrasive paper on a flat surface to ensure that the drilled faces are flat and free of burrs so that the power devices make good thermal contact.

Secure the fan to the appropriate end of the heatsink with four M3 × 25mm screws, making sure the arrow denoting the direction of airflow is pointing towards the heatsink. Orientate the fan so that the lead emerges at the corner shown in the photos. Now attach the finger guard together with its filter to the other end of the heatsink, using four M3 × 10mm machine screws.

Mount the heatsink assembly to the PCB with two M3 × 10mm screws with spring washers under the heads. The rectifier bridge (BR1) and the discharge Mosfet (Q7) can be mounted next, with a smear of thermal compound between the devices and heatsink to ensure good thermal contact. Use M3 × 10mm screws with spring washers under the heads. Don't solder the devices to the PCB just yet.

Next, mount the six IGBTs (Q1-Q6) after carefully bending their centre pins to fit the footprint. Again, use thermal compound, M3 × 10mm screws and spring washers.

Tighten all the devices down, making sure they don't twist too much, then solder and trim all the leads (of Q1-Q7 and BR1). Give all the screws a final tighten – you can't get to some of them once the DC bus capacitors are installed.

Affix the thermistor to the top of the heatsink, again using thermal compound, an M3 × 10mm screw and spring washer. Orient the thermistor lead along the heatsink towards the fan as shown. Trim and strip the thermistor and fan leads, then solder them to the PCB pads provided. The thermistor is not polarised, but the fan is, so make sure the red lead goes to the pad marked by the plus sign.

Now you can install all the remaining components. I suggest starting with the shortest and finishing with the five large electrolytic capacitors. Pay attention to the orientation of the filter capacitors – their positive leads must all go towards the top of the board!

Be careful also with the AC-DC power modules; they look similar but have different secondary voltages. The 15V one is MOD1 and the 12V one is MOD2.

You have finished the PCB assembly at this point, but it's a good idea to take a bit of time to check your work thoroughly before moving on.

## Enclosure preparation

The enclosure needs to have a square opening cut into the side to accommodate the heatsink exhaust, plus a series of ventilation holes in the top and opposite side and holes for the cable glands in the bottom end. The locations and dimensions of these are given in Fig.11.

Making the square opening can be a challenge. It helps to screw the lid firmly onto to the case for this operation, as the opening overlaps both the base and the lid. I applied masking tape in the area of the cutout and marked its edges onto that. I created the opening by chain-drilling a series of holes near, but just inside the marked line and then filing carefully up to it.

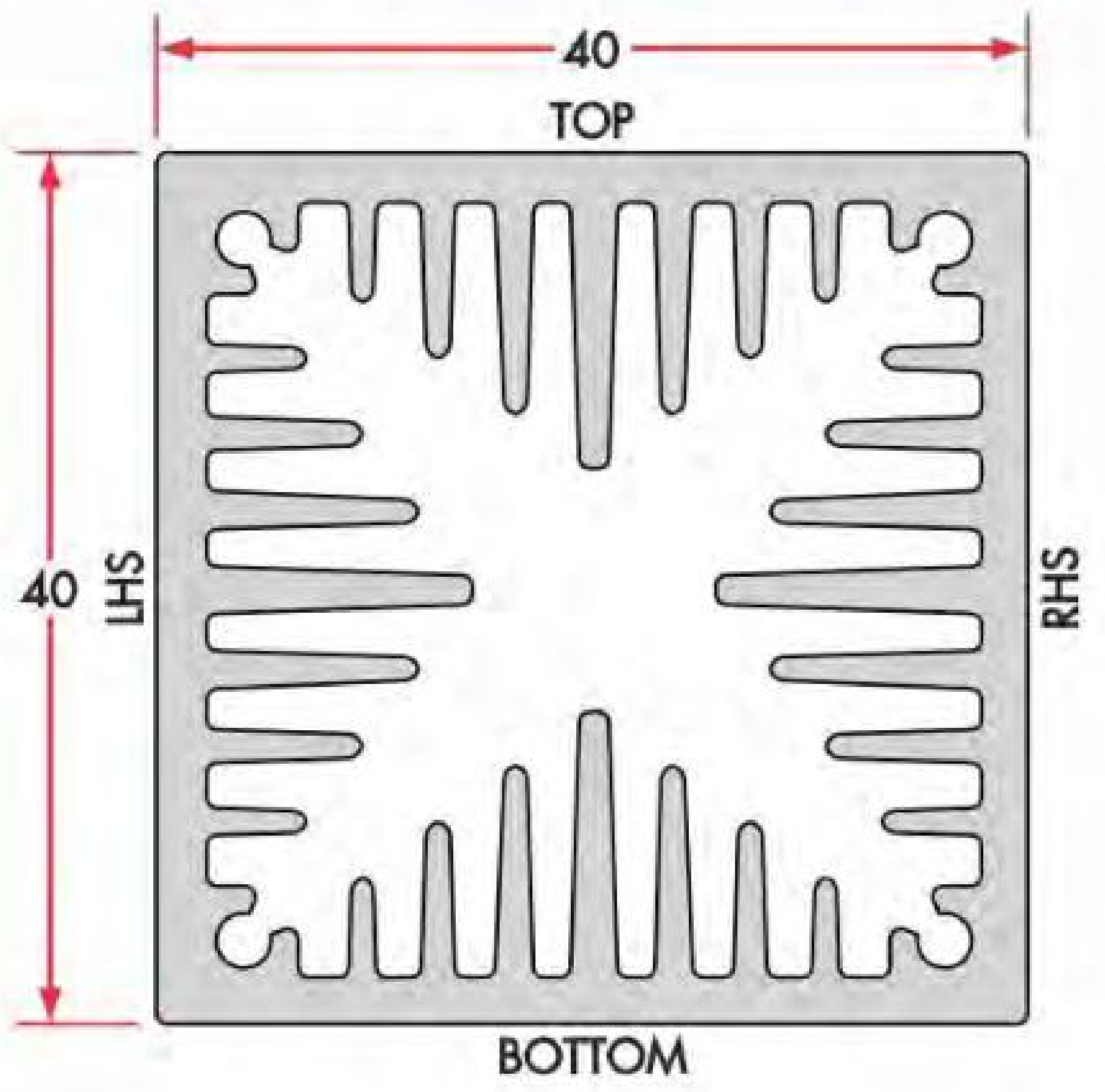
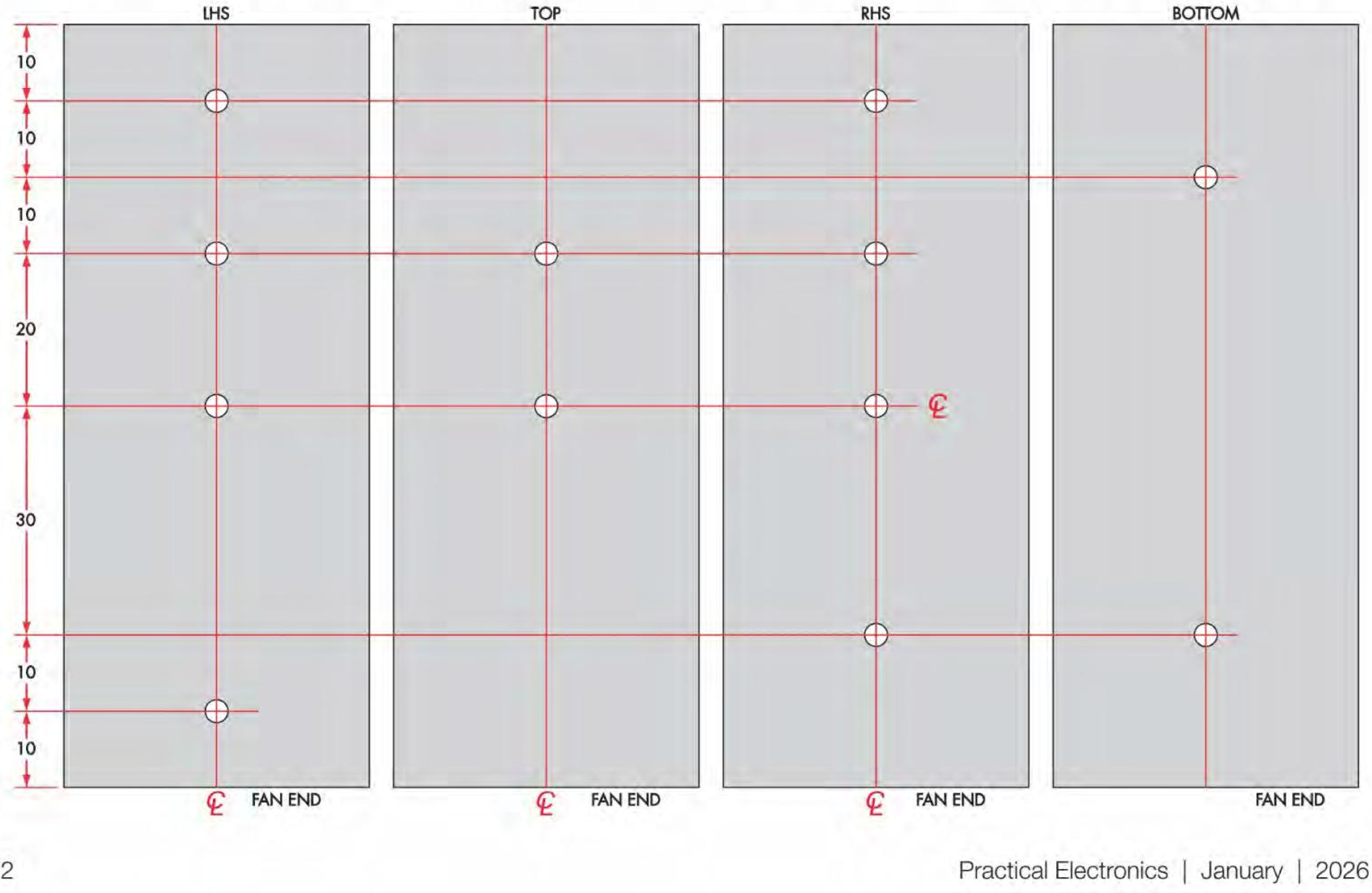


**Fig.10: the heatsink requires a total of 11 M3-tapped holes. They are positioned differently on each face, so be careful to get them all correct with respect to each other. All dimensions are in millimetres, and the diagram is shown at actual size.**

MATERIAL: ALUMINIUM TUNNEL HEATSINK EXTRUSION, 40×40×100mm ALL HOLES (11): M3.0×0.5mm TAPPED FOR 10mm



# Variable-Speed Drive Mk2, part two

Next, drill the 14 ventilation holes according to the diagram. I used masking tape as before to mark the centres, then drilled pilot holes with a 3mm drill bit, followed by a 10mm bit. You can then drill holes in the bottom end of the enclosure for the cable glands.

Two of the glands are required: one for the mains input and one for the motor output cable, but the third one, for control cables, is optional. If you are using the VSD in standalone mode (see the applications section below), this hole may be unnecessary. The hole size should match the glands that you use.

Make sure you get the correct sized glands for your cables – they will only provide good strain relief if they are matched to the cable diameter.

The enclosure comes with a length of O-ring material which you should push into the slot in the lid, avoiding the area of the fan guard cutout.

As a side note, you can get a set of mounting feet for the enclosure that allows it to be mounted on a panel or wall. If you are using those, now is a good time to screw them onto the bottom of the enclosure.

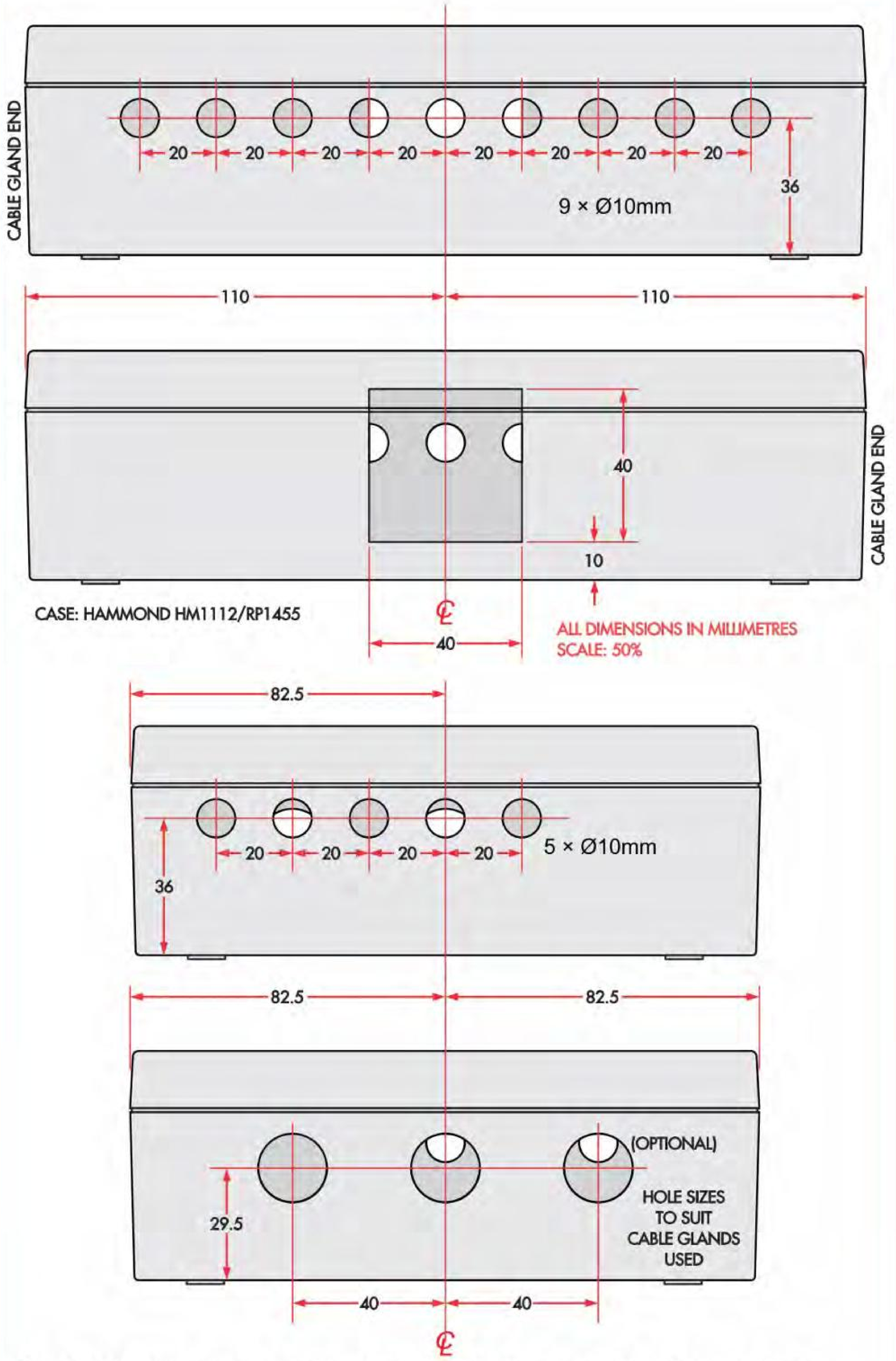


**Fig.11: the case needs a square opening for the heatsink exhaust, plus a total of 14 10mm ventilation holes as shown. The size of holes for the cable glands depends on the exact glands you are using.**

## Final assembly and wiring

You can now mount the PCB assembly into the case with four self-tapping screws and wire it up to suit your application. For most single-phase applications, an input cable with a three-pin mains plug and an output cable with a matching mains socket should work.

An easy way to create these cables is to sacrifice a low-cost extension cord by cutting it in half. Please use something that meets your local electrical standards, bought from a reputable supplier and not some random internet find.

Feed the cut end of each cable through the appropriate gland, tighten, and then crimp female 6.3mm spade connectors to the conductors. Either use insulated spade connectors for the Active and Neutral (brown and light blue) wires, or add some insulating heatshrink tubing in the appropriate colours over the exposed metal after crimping.

We need a direct 10A wire connection between the incoming and outgoing Earth wires to ensure the device can handle a high fault current if something goes wrong with the motor. Therefore, cut a 15cm-long piece of 10A green/yellow striped wire (which can be stripped from 10A mains flex or a spare 10A mains cord) and crimp piggyback spade lugs onto both ends.

Plug the incoming/outgoing Earth wire spades onto the tabs on the piggyback connectors and then shrink some 10mm green/yellow striped heatshrink tubing over the piggybacked connectors. They will be close to the Active and motor output spades. While those are also insulated, it doesn't hurt to have extra insulation!

Plug the piggyback spade lugs onto both Earth connectors on the PCB, then connect up the Active (brown), Neutral (blue) and motor output wires.

Double-check the wires are in the right places. The wire with the mains plug on the end (incoming power) must go to the A, EARTH and N spades near the fuse clips, while the one with the socket on the end goes to the EARTH, U and V motor connectors near IC2. Now is also a good time to insert the 10A slow-blow fuse into the F1 clips and slip the insulating cover over the top.

If you are driving a three-phase motor, or building the VSD into another piece of equipment, you may need custom wiring. In any case, it is absolutely mandatory to wire in the mains Earth and to connect the motor Earth to the motor chassis with a proper wire between the two (not relying on the PCB to conduct Earth current!).

The PCB Earth connections are for two purposes only: to Earth the heatsink for safety, and as part of the mains EMI filters that each have two Y2 capacitors between the phases and Earth. As mentioned in the adjacent caption, we recommend attaching the Earth wire directly to the heatsink as well.

## Control wiring

This speed controller has been designed to be as flexible as possible. In the standalone configuration, no external controls are required. The E-Stop and Run terminals should be bridged by short lengths of hookup wire, and the internal speed pot selected on S1.

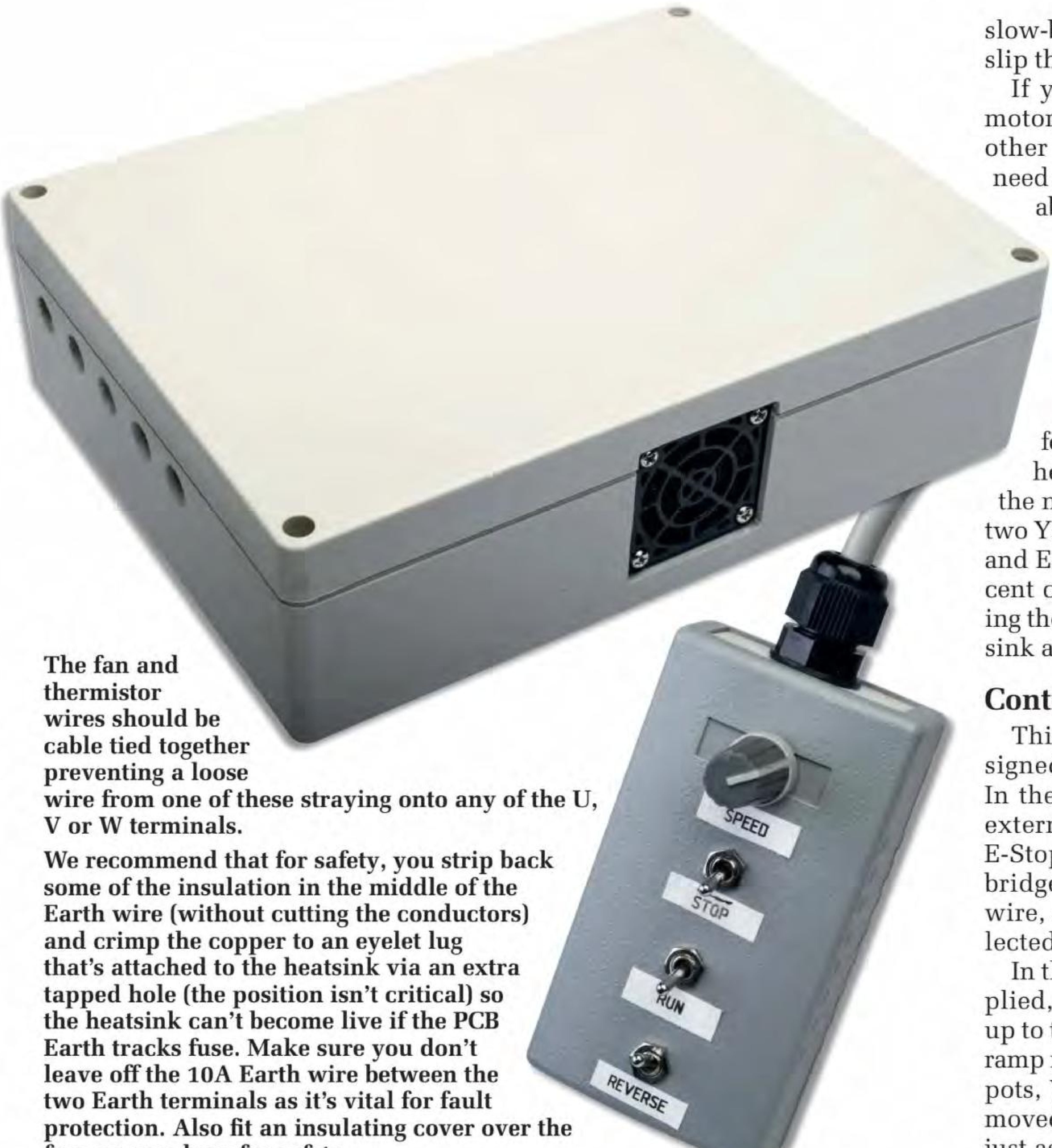
In this case, as soon as power is applied, the motor will start and ramp up to the preset speed. The speed and ramp rate are set via the onboard trimpots, VR1 & VR2. When power is removed, the motor will coast to a stop just as it would if switched off when directly connected to the mains.

This arrangement could be used to run a single-phase motor at a lower speed than usual, or to run a three-phase motor at a fixed speed from a single-phase supply. It could also be used as a 'soft starter', to provide a gentle start for sensitive loads or to limit the initial starting current surge. Most pool pump applications will also use this configuration.

At the other end of the spectrum, it is possible to use this controller as part of a more complex control system, such as for a machine tool. In such applications, the VSD would normally be mounted in an electrical cabinet, with external controls (run, emergency stop, speed control etc) located on a panel close to the operator.

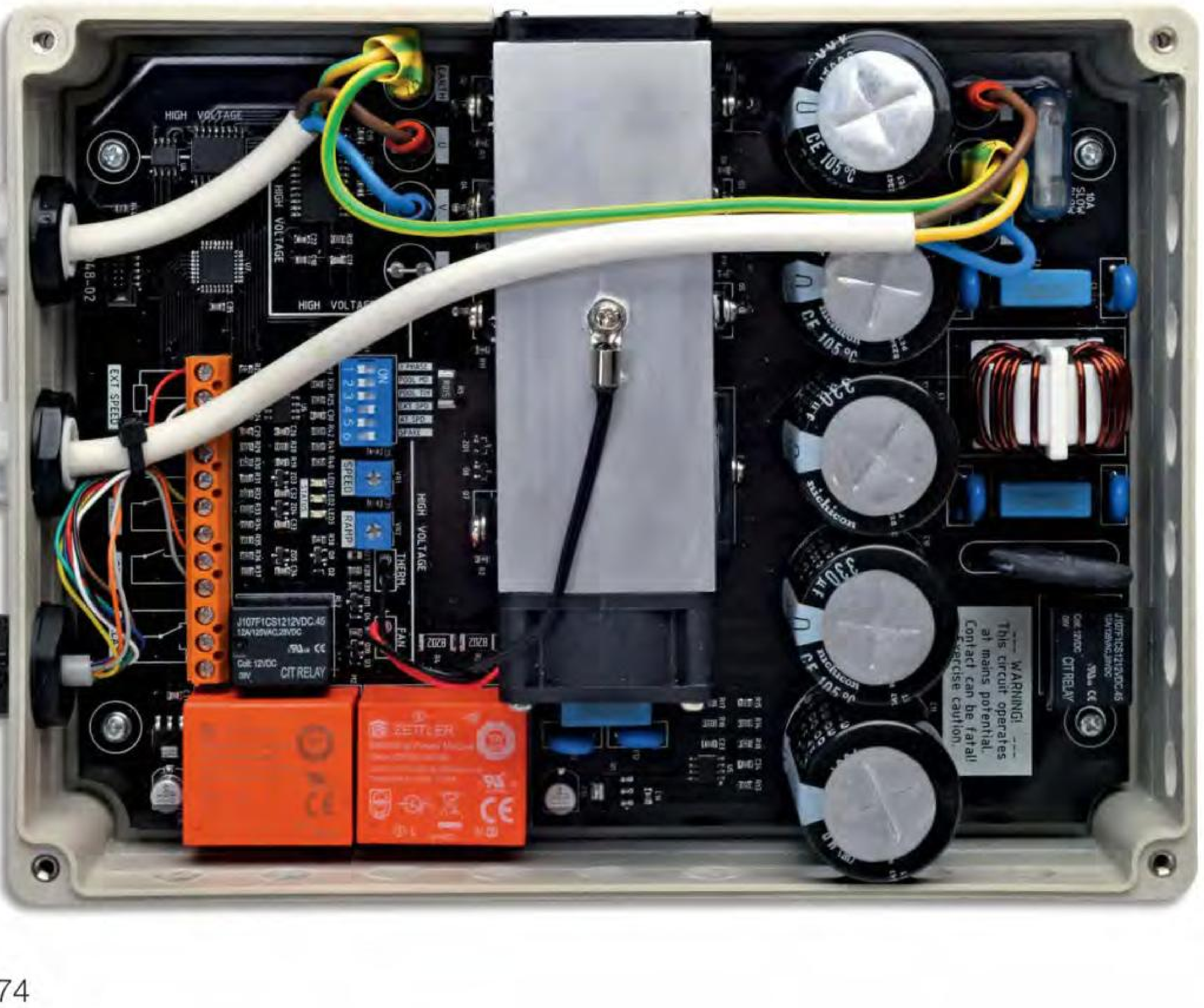
If the machine tool is numerically controlled, these control signals may come from a CNC controller or PLC.

You can see from our photos that



**The fan and thermistor wires should be cable tied together preventing a loose wire from one of these straying onto any of the U, V or W terminals.**

**We recommend that for safety, you strip back some of the insulation in the middle of the Earth wire (without cutting the conductors) and crimp the copper to an eyelet lug that's attached to the heatsink via an extra tapped hole (the position isn't critical) so the heatsink can't become live if the PCB Earth tracks fuse. Make sure you don't leave off the 10A Earth wire between the two Earth terminals as it's vital for fault protection. Also fit an insulating cover over the fuse as seen here for safety.**


HIGH VOLTAGE
EXT SPEED
SPEED
RAMP
THERM
FAN
WARNING
This circuit operates at mains potential. Contact can be fatal! Exercise caution.
ZETTLER
CIT RELAY

we built a small ‘remote control’ box to test out the external control functions. It’s little more than three switches and a pot mounted to a Jiffy box and wired to a 9-core alarm cable, run through cable glands into the VSD case, where they connect to the EXT SPEED, ESTOP, RUN and REV terminals of CON8-CON10.

We won’t go into details here, as we expect anyone who can build this VSD will be able to figure out the wiring from the PCB labelling.

The cable gland outside nuts that are tightened to secure the mains input and output wires should be permanently fixed using super glue on the threads to prevent the glands from being undone from outside the box and the mains wires becoming loose.

## Using the VSD

Using the VSD is straightforward. If the unit trips out when starting, you can extend the ramp rate and/or switch the BOOST DIP switch on. We tested it on a domestic pool pump and found that, with the correct settings, it had no trouble starting the pump under load. If you have one, you can use a current clamp meter around one of the motor power wires to monitor the motor current during startup.

The VSD should be able to deliver its full rated current (9A in single-phase mode and 5.5A in three-phase mode) continuously and up to 18A/11A for a few cycles. You will need a clamp meter with a peak hold setting to measure this.

If you are wiring the VSD directly to the motor, you will need to work out how to connect it. Single-phase PSC motors have notoriously confusing terminal housings with no discernible standard arrangement. There is usually a diagram inside of the terminal housing lid to help; otherwise, see if you can locate a wiring diagram for your motor online. Don’t forget to connect the Earth wire solidly to the stud provided in the terminal box.

The only way to change the direction of rotation of PSC motors is to reverse the sense of the start winding with respect to the run winding. Many motors have an arrangement of relocatable bridges to allow this to be done without rewiring the whole motor.

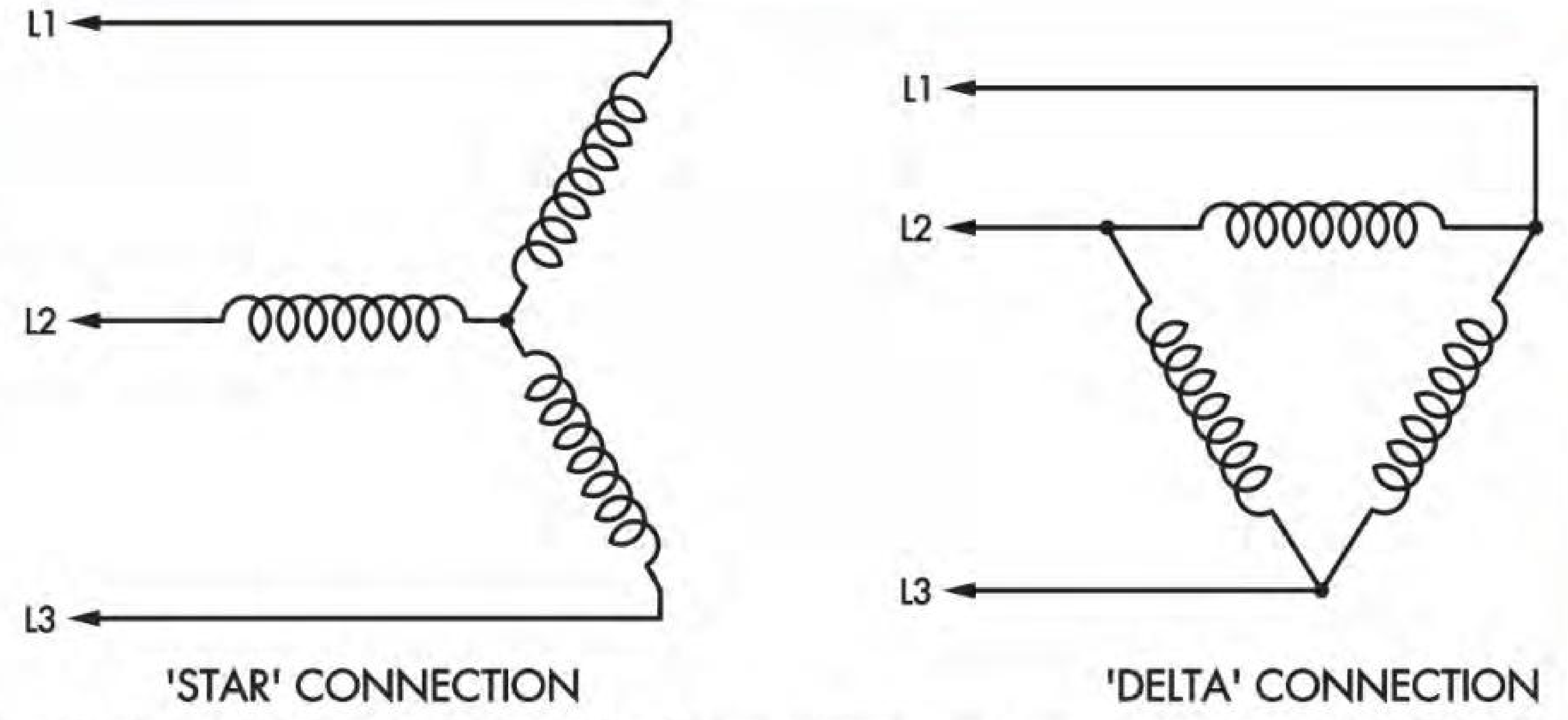


**Fig.12: the windings of small 3-phase motors are normally connected in star configuration for use with the 400V RMS 3-phase mains supply. In this case, each winding is driven with the phase-to-neutral voltage of 230V. By changing how the windings are connected (which can usually be done by moving some jumpers), the motor can be changed to delta configuration, with just one winding between each phase. It can then be driven from a 230V RMS 3-phase supply such as the output of this motor controller.**

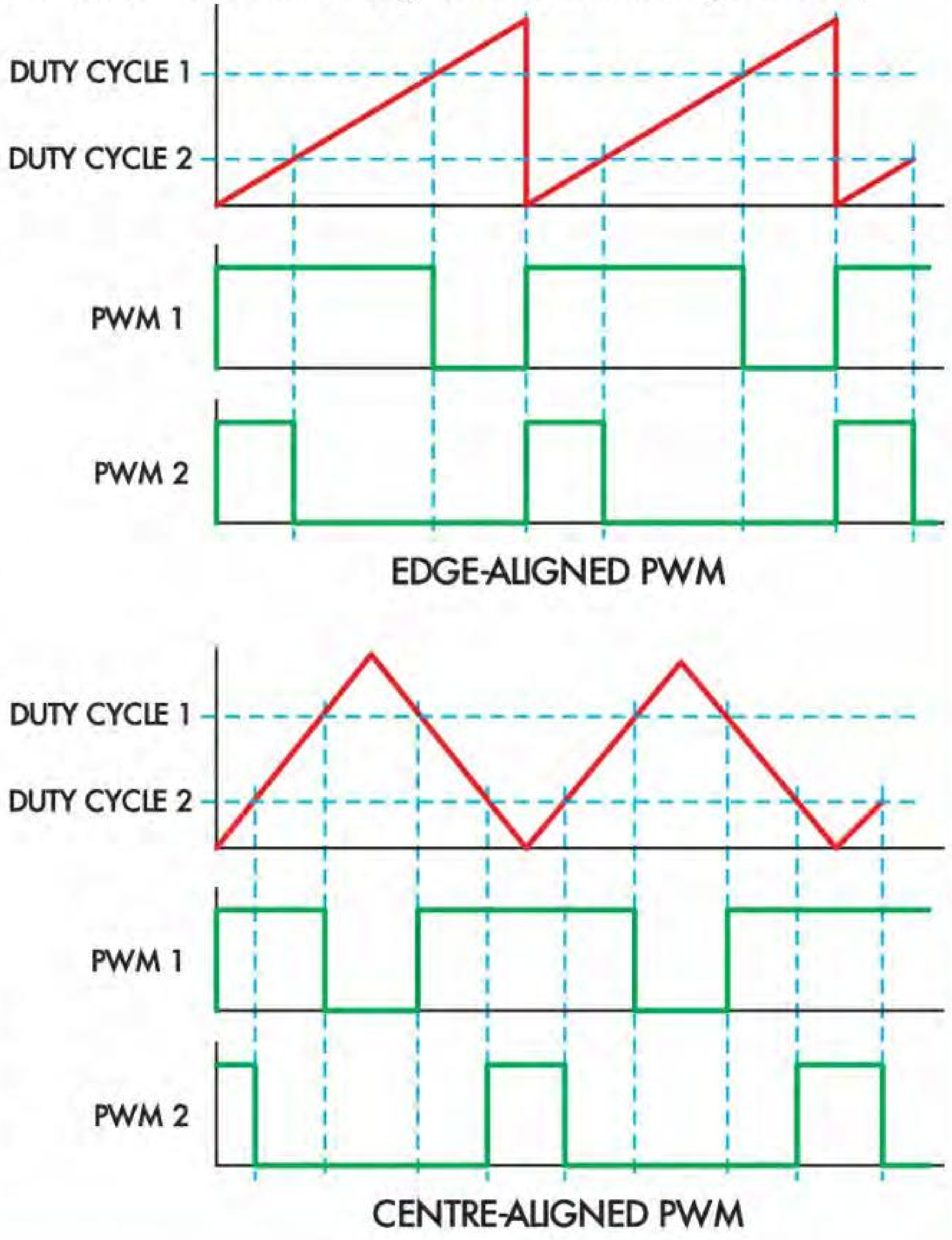


**Fig.13: this diagram illustrates the difference between traditional edge-aligned PWM and centre-aligned PWM (also known as dual-ramp PWM). With centre-aligned PWM, the leading edge of each pulse moves as the duty cycle changes. This is an advantage because if all outputs switch high at the same time, as with edge-aligned PWM, the total current pulse is larger and so more EMI is generated.**

The terminal arrangement for three-phase motors is usually a little simpler. The VSD can only supply a phase-to-phase voltage of 230V RMS, so it is suitable for motors with 230V or 240V windings (most small induction motors). The rating plate will normally quote the voltage rating as 230V/400V, 240V/415V or something similar.

There are usually six terminals for the three windings, with bridges to connect the windings in star (Y) configuration for the higher voltage or delta (Δ) configuration for the lower (see Fig.12). For 230/240V operation, use the delta (Δ) option. Again, the inside of the terminal box lid should have a diagram to help.

You can connect the VSD’s U, V & W outputs in any order, although this will affect the direction of rotation. If the direction is not what you want, swap any two of the leads or use the Reverse control input, which does the same thing electronically. Again, connecting the Earth is mandatory for safety.

A word of warning: induction motors often have a shaft-mounted fan that blows cooling air across the fins cast into the housing. This fan will be much less effective at low shaft speeds, so be careful if you intend to run a motor in this way for long periods of time or in very hot environments. If this is a concern for you, consider using an external cooling fan with a separate power source. PE

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PCBs for most recent *PE/EPE* constructional projects are available. From the July 2013 issue onwards, PCBs with eight-digit codes have silk screen overlays and, where applicable, are double-sided, have plated-through holes, and solder mask. They are similar to photos in the project articles. Earlier PCBs are likely to be more basic and may not include silk screen overlay, be single-sided, lack plated-through holes and solder mask.

Always check price and availability in the latest issue or online. A large number of older boards are listed for ordering on our website.

**In most cases we do not supply kits or components for our projects. For older projects it is important to check the availability of all components before purchasing PCBs.**

Magazine back issues are available – see **https://www.electronpublishing.com/product-category/pe-back-issues-print-and-digital/**

# Double-sided | plated-through holes | solder mask

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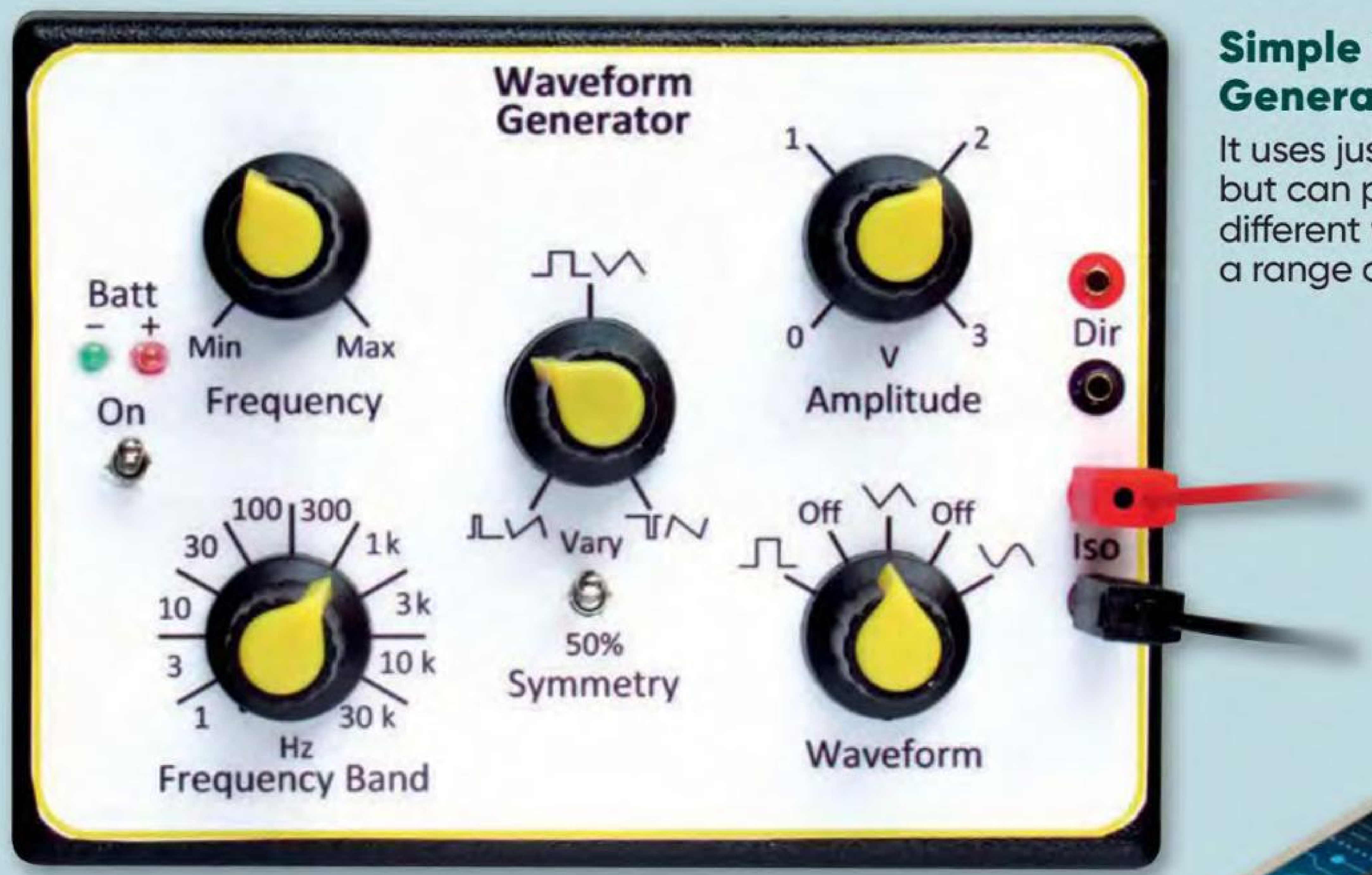

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